SOLID-STATE DIGITAL IN-SITU
ACOUSTIC DATA ACQUISITION

David Michael Craig

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THESIS

SOLID-STATE DIGITAL IN-SITU
ACOUSTIC DATA ACQUISITION

by

David Michael Craig

December 1975

Thesis Advisor: V. M. Powers

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and on methods for effecting large-scale data reductions. Recent advances in integrated-circuit electronics are evaluated in terms of suitability for use in instrumentation requiring both large-scale memory capacity and low power consumption. Results of the research include functional block diagrams of an ambient sea-noise recorder and a shipping-container impact recorder.

Solid-State Digital In-Situ

Acoustic Data Acquisition

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ABSTRACT

The problems of acoustic environmental data recording are studied to develop an alternative to conventional magnetic-tape digital storage for <u>in situ</u> recording instruments. Various physical data transducers are analyzed to determine which types are best suited for portable solid-state environmental recorders, and a discussion of information processing concentrates on the problems of high-density versus low-density digital data storage and on methods for effecting large-scale data reductions. Recent advances in integrated-circuit electronics are evaluated in terms of suitability for use in instrumentation requiring both large-scale memory capacity and low power consumption. Results of the research include functional block diagrams of an ambient sea-noise recorder and a shipping-container impact recorder.

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I. INTRODUCTION

Data gathering by measurement and recording *in situ* is an old and accepted branch of acoustics since few acoustic quantities are suitable for laboratory measurement. Until the recent past, the choice of recording instruments has been somewhat limited. Most devices which measure and record shock tend to be clockwork- or battery-powered strip-chart recorders, while ambient noise recordings, in either analog or digital form, are invariably accomplished using battery-powered magnetic-tape drives.

The objective of this research is to investigate how state-of-the-art solid-state electronics technology can be applied to the problems of measuring and recording of acoustic and related data. In order to illustrate specific techniques two particular applications will be attempted:

A. Design an instrument which will measure and store a time record of ambient sea noise.

B. Design an instrument which will measure and record the dynamic history of a shipping container exposed to a shock environment during transportation.

It is expected that these instruments will be completely self-contained and capable of functioning for a long period without external control. Both will take advantage of microprocessor capabilities to achieve reduction of the

input data from its low-density analog form to some high-density digital form. It is asserted that if these two recorders are shown to be feasible, then similar devices could be constructed for many different environmental recording uses.

In simplest form, either of the proposed recording instruments could be described as consisting of transducers which sense the physical quantities; a data management system which performs signal-conditioning operations that may include amplification, filtering, analog-to-digital conversion, and mathematical manipulation of data; and a memory device for ultimate storage of the data.

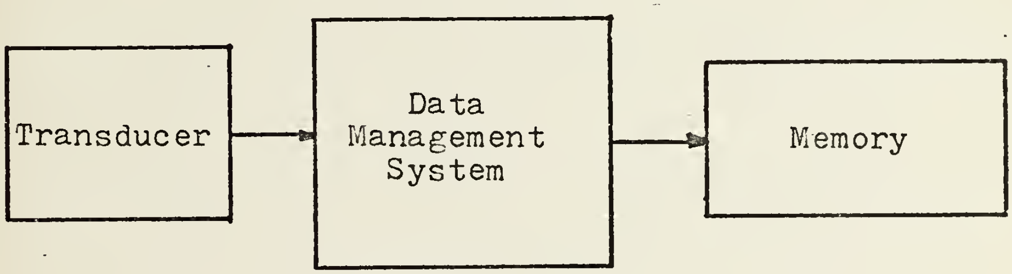


Figure 1. A simplified recorder diagram.

Toward achieving the stated design goals, attention will first be given to the physical quantities of interest with emphasis on selecting suitable transducers for each. Next, signal processing of the analog data will be considered and emphasis will be placed on techniques for minimizing the amount of data which must be stored. Subsequently,

the state of the art in integrated-circuit technology will be explored to select that technology which will be most efficient for portable instrument design. Last, designs will be presented for both an ambient sea-noise recorder and a shipping container environmental recorder which will employ those concepts developed in this research.

II. PHYSICAL QUANTITIES AND TRANSDUCERS

This chapter will describe the principal physical quantities which are measured and recorded in situ. The transducers required to sense each quantity will be discussed in sufficient detail to permit selection of those most suitable for a portable instrument. Simple models of the data structure will be used when appropriate to aid in selection of the transducer.

The physical quantities of principal interest are ambient sea noise and transportation shock. Discussions of temperature and relative humidity will be included since these are of interest when monitoring the dynamic history of a shipping container.

Solutions to the problems associated with the recording of these physical quantities are expected to be closely related to solutions of problems associated with other physical quantities such as salinity, hydrostatic pressure, or seismic vibration which have similar analog data structure.

A. AMBIENT NOISE AND HYDROPHONES

1. Introduction

Acoustic signals in the 0-20 kHz spectrum can vary from the complexity of classical music or human speech to the relative simplicity of ambient noise. Recording techniques for the many possible signals vary according to the

desired results. Usually, exact reproduction of these signals is desired; but often, only knowledge of one characteristic of the signal is desired. Such signals afford the opportunity for signal conditioning which can greatly reduce the amount of data which must be stored in the recorder. One such signal of particular significance is ambient ocean noise. Numerous systems have been devised for recording this phenomenon, but these invariably have been based upon large, broad-band magnetic tape recorders. This discussion will address the problems of recording ambient noise with a view toward designing a solid-state memory. Many of the points considered will apply equally well to the recording of other acoustic phenomena.

2. Ambient Noise

Ambient ocean noise is defined as that portion of the total noise background observed with a non-directional hydrophone which is not due to the hydrophone sensor or its mounting nor to any other readily identifiable noise source. Validity of ambient noise measurement requires that (1) all possible sources of self-noise be eliminated or reduced to an insignificant level (e.g., cable strumming, splashing waves against hydrophone cables, 60-Hertz hum, or crabs crawling on a bottom-mounted hydrophone), and (2) all identifiable distant sources, such as individual ships, must not contribute to the noise background.

Ambient noise measurements have been made over a frequency range extending from below 1 Hertz to about

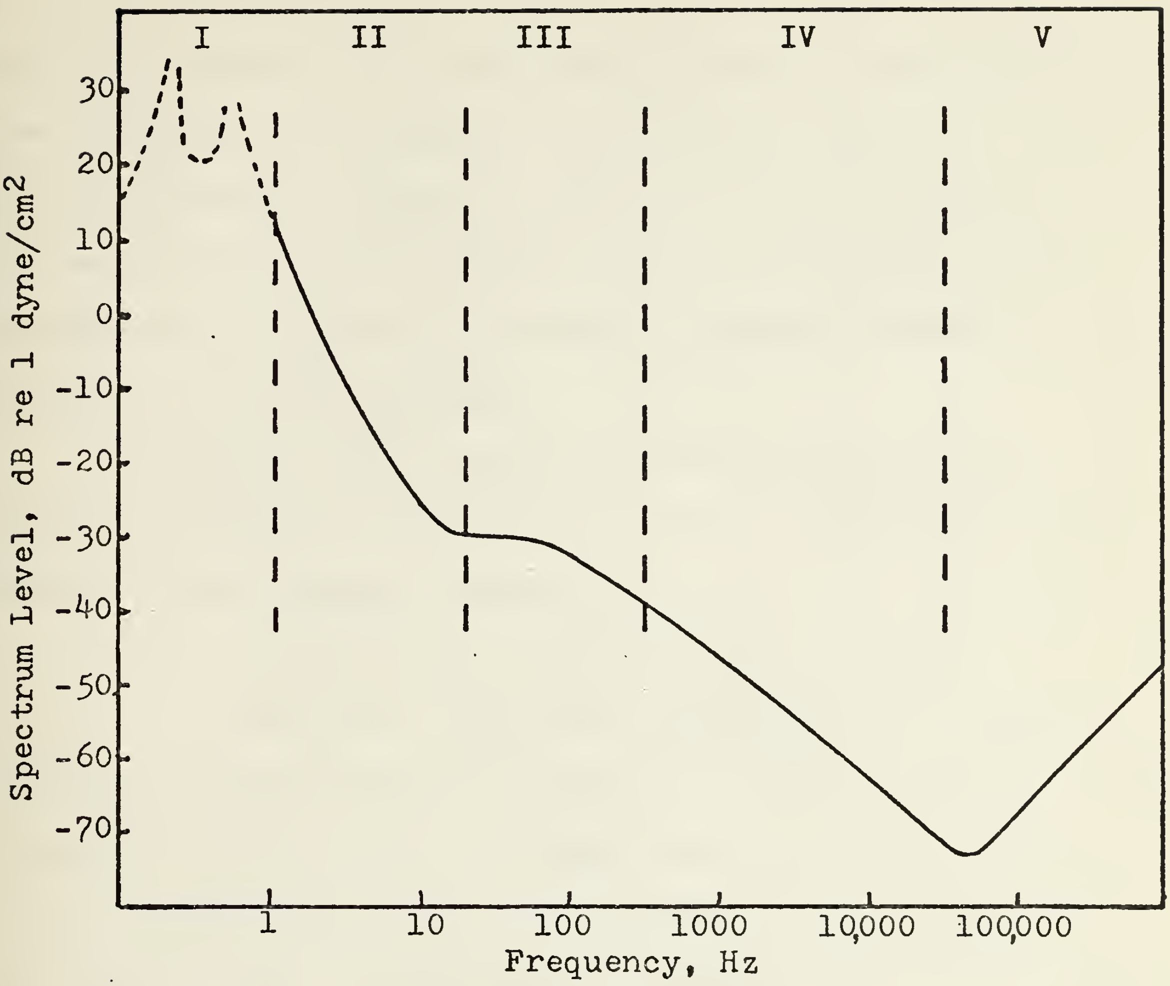


Figure 2. Ambient deep-sea-noise spectrum.

100 Kilohertz. The data over this range shows different characteristics in different portions of the band. A composite spectrum of deep-ocean noise has been described by Wenz which shows five significantly different portions of the spectrum. Each of these portions is dominated by one or more types of noise source.

The sources of ambient noise in Band I (< 1 Hz) are largely unknown but are believed to be primarily of seismic or hydrostatic pressure origins. Band II (1-20 Hz) effects

are due primarily to ocean turbulence in deep water, but some noise in this band is due to surface wind effects. Band III (20-500 Hz) is dominated by distant ship traffic. The "plateau" which occurs in this band is shown by nearly all measurements. Band IV (500-50 kHz) noise originates at the sea surface close to the location of the measurement hydrophone. In Band V (> 50 kHz), increasing thermal noise dominates all other sources.

Urick [40] has concisely described each of the sources of ambient noise occurring in the bands defined by Wenz. A brief summary follows.

a. Tides and Hydrostatic Effects of Waves

Since most underwater hydrophones are pressure sensitive, they respond to changes in ambient pressure whether of acoustic origin or not. Thus tides and surface waves can cause hydrostatic pressure changes of large amplitude but usually very low frequency. Surface wave effects diminish rapidly with depth but can be dominant in the low-frequency background of a hydrophone located in shallow water. Tidal effects are generally on the order of 1 to 2 cycles per day and therefore usually below the frequencies of interest.

b. Seismic Disturbances

A strong and nearly continuous form of seismic activity called <u>microseisms</u> having regular periodicity of about 1/7 Hertz and a vertical amplitude on land of about

10^{-4} centimeters are believed to be a major source of sea noise at frequencies below 1 Hertz provided the same vertical amplitude exists in the deep sea bed. Volcanos also provide intermittent noise in this frequency band.

c. Ocean Turbulence

Irregular or random water currents of larger or small amplitude can create noise by several mechanisms. They can shake the hydrophone or its mounting, turbulent pressure changes can be radiated and appear as background noise originating at a distance, or most importantly, pressure changes inside the turbulent region can be picked up by a pressure-sensitive hydrophone. A contribution in the 1-10 Hertz spectrum band has been inferred.

d. Ship Traffic

Evidence suggests that distant ship traffic is a principal source of noise at frequencies in the decade 50-500 Hertz and the dominant source about 100 Hertz.

e. Surface Waves

Noise level at the measurement hydrophone has been strongly correlated to local wind speed which couples to the sea as underwater sound through the mechanism of surface waves. The exact method of coupling is not yet known. The dominant effect of surface waves is in the 1-30 kiloHertz band.

f. Thermal Noise

Thermal noise due to the molecular motion of seawater becomes dominant about 100 Hertz.

g. Intermittent Sources of Ambient Noise

Intermittent sources of ambient noise are primarily rain or biologics. Sea creatures may be of special concern in shallow water.

Deep-water ambient noise generally conforms to Wenz's predicted curves. Shallow-water noise is not as well defined. Wide variations are possible between different locations or at different times. Studies have shown that the noise level in bays and harbors exceeds that of deep water particularly at frequencies below 100 Hertz. In coastal waters wind speeds play a major role in determining the noise levels at frequencies between 10 and 3000 Hertz.

Ambient noise levels show some variability due to intermittent sources or changes in dominant sources. These transients may account for 5-10 dB variance from Wenz's predicted levels. In general, the amplitude distribution of ambient noise is gaussian at moderate depths but spiky near the surface or under ice. Under-ice noise tends to be discontinuous due to tensile cracking and rubbing of ice masses which dominate the noise intermittently. Noise levels under ice are 5-10 dB higher than in open water and present their own special problems in terms of noise level recording.

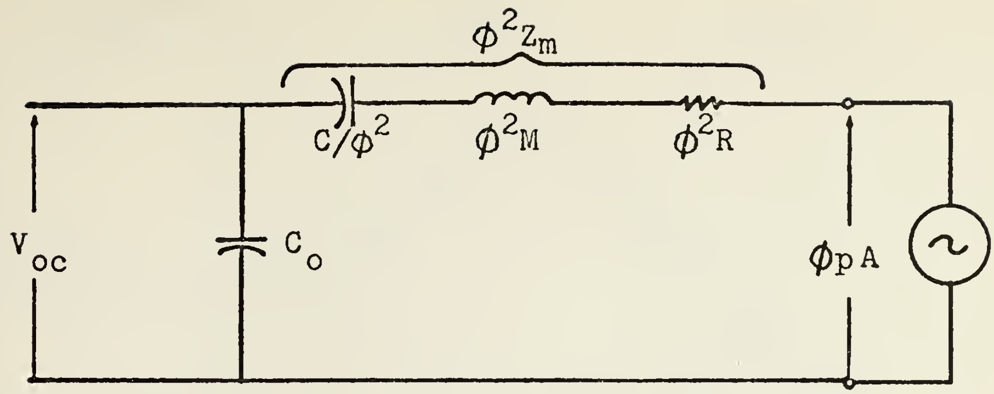
The characteristics of the ambient noise in the band of interest must be fully considered when designing a system for recording that noise. Each spectrum band offers its own unique problems for recording and signal processing.

3. Hydrophones

Measurement transducers for ambient noise include both standard hydrophones and other specialized transducers. The chief requirements of a standard hydrophone are linearity and stability. The sensitivity of an ideal standard hydrophone should be independent of time, frequency, and environment--particularly pressure and temperature. Special measurement transducers are those designed for special purposes. For example, the pressure-gradient hydrophone has a directivity pattern which may be especially useful for some measurements but undesirable for general use. Frequently, a resonant transducer is used when high sensitivity is desired although wide bandwidth is sacrificed. A third, but very important, class of special transducers are those which conform to the electroacoustic reciprocity theorem [4].

The most commonly used measurement transducers are of the piezoelectric crystal or ceramic, magnetostrictive, and moving-coil types. Other types are sometimes used, but rarely, and will not be considered here.

The standard hydrophone has several basic characteristics which must be considered when designing a system for recording ambient noise. A hydrophone used for noise measurement requires very wide band response. Often a useful frequency range of several decades is required for such purposes as measurement of broadband ambient noise. The useful frequency range is determined by several transducer parameters:



Z_m = mechanical impedance, a function of frequency; ϕ = electromechanical voltage/force conversion factor; R = mechanical resistance; M = effective mass; C_o = blocked electrical capacitance; C = short circuit compliance; p = acoustic pressure; A = diaphragm area.

Figure 3. Electromechanical equivalent circuit for the piezoelectric hydrophone.

a. The sensitivity

b. The electrical impedance

c. The transducer's mechanical limitations

d. The contour of the sensitivity-versus-frequency curve

Limitations (a) and (b) are related and are determined primarily by the acoustic and electrical noise levels encountered during measurement. These parameters are flexible, and no useful limits can be easily established for them. Limitation (d) can be examined in detail since the frequency response can be predicted theoretically. However, in actual use, spurious resonances and other effects may occur which alter the predicted behavior [4].

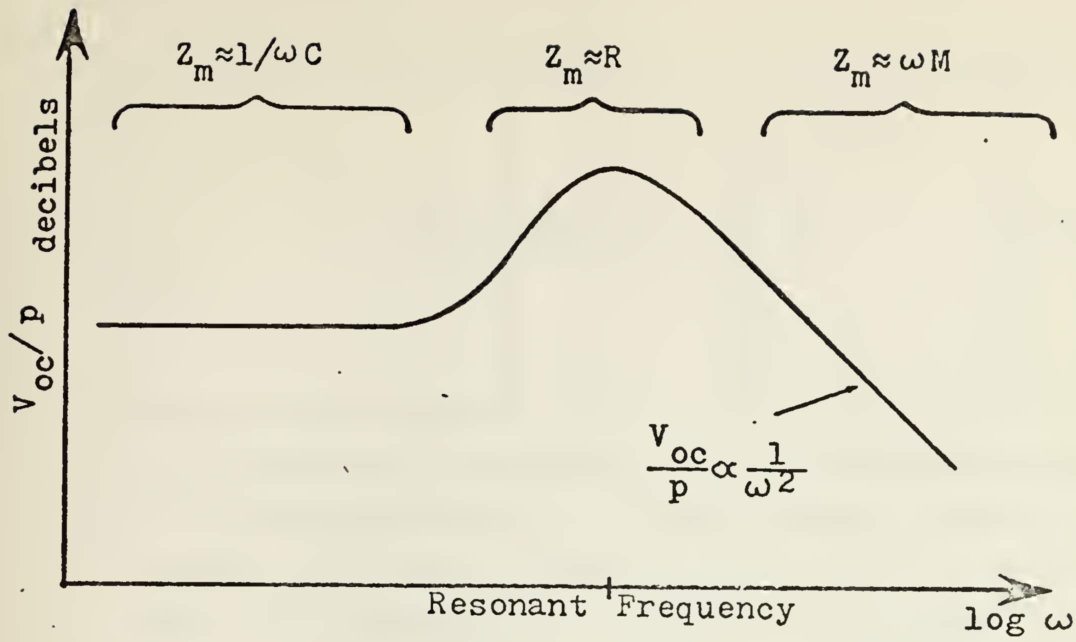
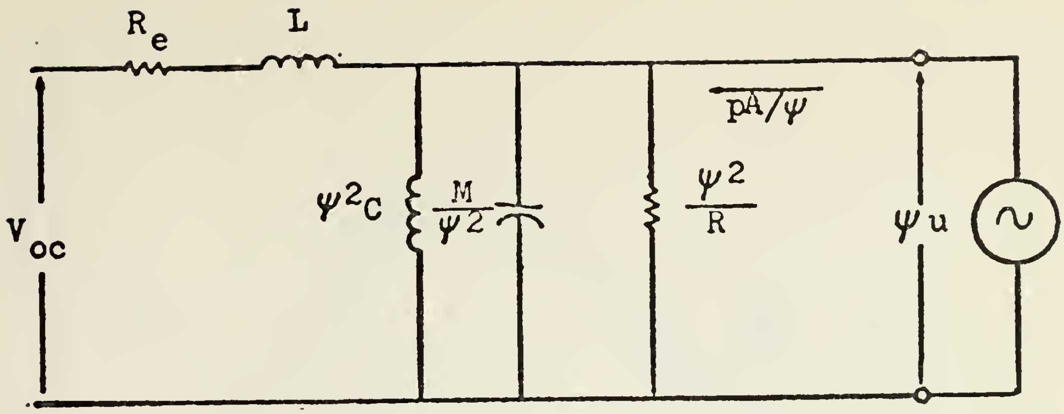


Figure 4. Sensitivity-versus-frequency for the piezoelectric hydrophone.

The piezoelectric transducer may be described in analog form by the equivalent circuit shown in Figure 3 [4,7]. An expression for the sensitivity of a piezoelectric hydrophone can be written in terms of the equivalent circuit elements:

$$\frac{V_{oc}}{p} = \frac{\phi A}{j\omega C_o \phi^2 Z_m + 1} \qquad (1)$$

At all frequencies except near the resonant frequency $j\omega C_o \phi^2 Z_m >> 1$. Near resonance, $1/j\omega C \simeq j\omega M$, $Z_m \approx R$, and the sensitivity peaks. At low frequencies, $1/j\omega C >> R + j\omega M$ and the sensitivity is approximately constant. At frequencies above resonance, $j\omega M >> R + 1/j\omega C$ and the sensitivity



R_e = electrical resistance; L = blocked inductance; ψ = electromechanical voltage/velocity conversion factor; C = open-circuit compliance; M = effective mass; R = mechanical resistance; p = acoustic pressure; A = diaphragm area; u = diaphragm velocity; V_{oc} = open-circuit voltage.

Figure 5. Magnetostrictive hydrophone electromechanical equivalent circuit.

decreases inversely with the square of the frequency and the response also goes through a series of anti-resonances which cause the hydrophone to be unsuitable for use as a measurement standard at frequencies above the first resonance. The behavior described is shown in Figure 4.

The analog circuit for a magnetostrictive hydrophone follows similar development but is unlike that of the piezoelectric transducer due to the different transduction principles involved [4,7]. The sensitivity of the transducer is given by the relation

$$\frac{V_{oc}}{p} = \frac{\psi A}{\frac{1}{j\omega C} + j\omega M + R} = \frac{\psi A}{Z_m} \qquad (2)$$

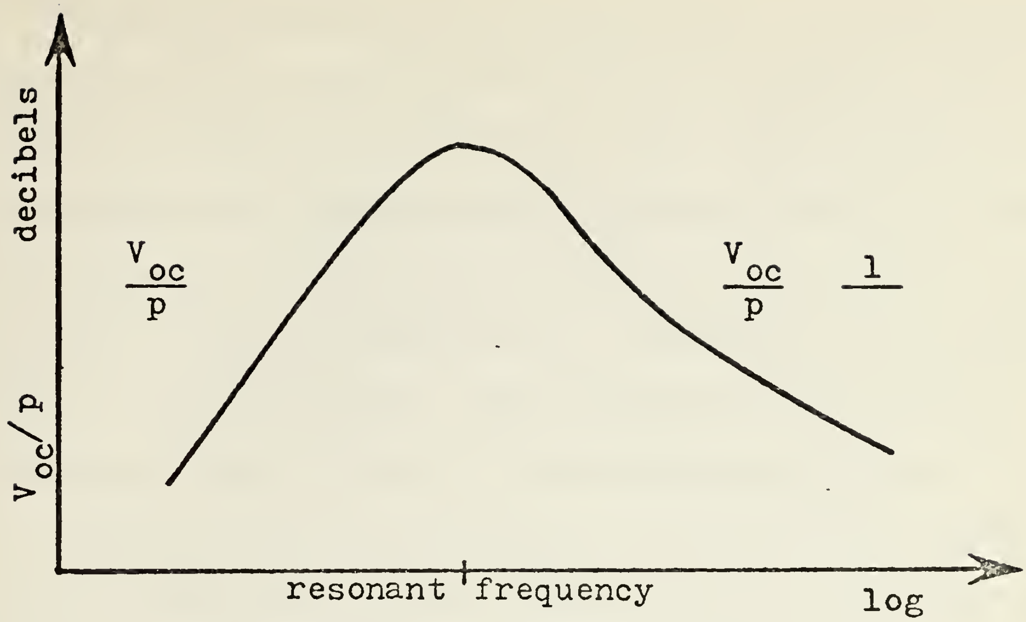


Figure 6. Sensitivity-versus-frequency of the magnetostrictive or moving-coil hydrophone.

That is, the sensitivity of the moving-coil or magnetostrictive transducer is inversely proportion to the mechanical impedance. The sensitivity is plotted in Figure 6 as a function of frequency.

It can clearly be seen that the best way to obtain the frequency-independent sensitivity desirable in a measurement hydrophone is to use a piezoelectric transducer operating well below its resonant frequency. Predictably, a wide range piezoelectric transducer would be one with a high resonant frequency since it has uniform response below resonance.

Underwater piezoelectric transducers are manufactured from piezoelectric crystals such as quartz, Rochelle salt, ADP (ammonium dihydrogen phosphate), and lithium sulfate, or from ferroelectric ceramics such as barium titanate and lead zirconate titanate. The ceramics have largely replaced the crystals due to ease of fabrication, ruggedness, the variety of special shapes available, high dielectric and piezoelectric constants, and low cost. The performance of the ceramics, however, is not as predictable as that of the crystals so that the latter may be more desirable when great stability is required under varying environmental conditions.

B. SHOCK AND ACCELEROMETERS

1. Introduction

This section will consider the vibration phenomenon of shock in view of selecting appropriate devices for measurement and recording *in situ*. Shock and impact are basically the same concept and refer to the transient transmission of mechanical energy to or from a system. In this discussion the terms will be used interchangeably. This discussion will be further restricted to the consideration of shock in transit situations. In particular, the dynamic behavior of a shipping container and methods for measuring such behavior and recording its history will be considered.

Damage during transportation can be attributed to several sources, but the most significant is improper handling. Impacts can occur due to rail-car switching or collisions,

vehicle braking, road surface effects, dropping of packages during handling, or from other sources. Impacts resulting from dropping tend to be greater in magnitude than those resulting from other causes.

A package experiences both impact and vibration due to vehicle motion during transit. It is seldom possible to distinguish between the two contributions. Truck shipment is the roughest mode of transportation with typical vibrations on the order of 1g and maxima of 4g. Rail and aircraft are intermediate with vibrations of 0.5g-2g, and ships provide the smoothest ride with vibrations typically averaging about 0.25g-1g [3].

Packaging greatly influences the maximum acceleration experienced by the contents of a shipping container, but because the fragility of items can only be estimated in most cases, package design is more an art than an exact science. Transportation shock has not been studied as intensively as some other aspects of mechanical vibration and many of its problems remain controversial. Measuring and recording instruments can collect analytic data for research which makes suitable packaging design possible.

Measurement of impact in transit actually can have several objectives: allocation of responsibility for damage, indication of required vehicle maintenance, classification of the ride, monitoring the ride, and research. Numerous instruments have been devised for impact measurements, but these have been predominantly mechanical in nature thus

limited in capabilities. All of these instruments can be classified into four types according to function. These types are listed in Table I [3].

TABLE I. Impact-measuring instrument classification

	Function	Question Answered by Instrument
a.	Indicate,Classify	Did at least one impact exceed a particular value?
b.	Measure,record	What was the greatest impact?
c.	Measure,record	What was the number of impacts above a particular value and the value of each impact?
d.	Measure,record	Same as above together with the time of occurrence of each impact.

Using LSI solid-state technology, it should be possible to implement any or all of these functions in one instrument or incorporate new functions into the measuring device.

2. Mechanical Transient Vibration of the Simple System

The response of a single-degree-of-freedom mechanical system to various input transients will be studied to establish what parameters must be measured and recorded to best describe the impact response of a shipping container.

The following discussion will follow the developments of Snowdon [27] and Macduff and Curreri [28] who describe the behavior of a resiliently mounted piece of apparatus (an accelerometer), the foundation of which experiences a transient displacement. The foundation is assumed to be the housing of the accelerometer rigidly attached to a recording

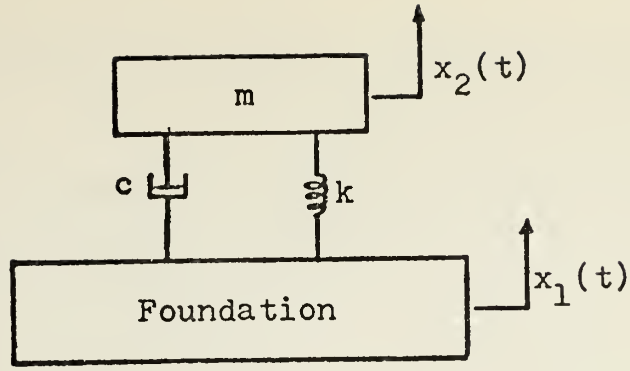


Figure 7. Transiently-disturbed simple system. An idealized seismic instrument.

instrument which is rigidly attached to a shipping container or other object whose dynamic behavior is to be studied. The theoretical response of such a system is discussed in Ref. 37 in terms of a rounded displacement step and a rounded displacement pulse input transient. The rounded transients are used in recognition of the fact that a rectangular pulse is not physically attainable. Both the steplike and pulselike displacements have finite first and second derivatives for all values of time, t. References 21 and 28 base their developments upon a true rectangular step displacement (while noting that it cannot be realized physically), as well as upon a sine-wave pulse input transient.

The system can be modeled as shown in Figure 7 where the system is represented by a mass, m, attached to a foundation by a spring of stiffness, k, and a dashpot with a viscous-damping coefficient, c. The undamped natural frequency of this system is given by

$$\omega_o = \sqrt{\frac{k}{m}} \text{ radians per second} \quad (3)$$

The coefficient of viscosity that corresponds to "critical damping", that is, which corresponds to an exponentially damped solution to the equation of motion which is just non-oscillatory, is given by

$$c_c = 2m\,\omega_o = 2\sqrt{km} \qquad (4)$$

Critical damping results in the displaced system being returned to rest in the most rapid manner. The damping ratio (or fraction of critical damping), $\delta_R = c/c_c$, is of considerable interest in the design of seismic instruments.

The equation of motion for the system in Figure 7 is given by

$$m \frac{d^2x_2}{dt^2} + (k + c\frac{d}{dt})(x_2 - x_1) = 0 \qquad (5)$$

Solutions to the equation of motion for displacement, velocity, or acceleration are obtained in a straightforward manner by assuming displacement input transients, x, of one of the forms discussed. However, the exact form of the solution can be quite complex and is most easily understood by examining generalized response curves.

Figure 8 shows the acceleration response of the system to a rectangular pulse of acceleration of duration τ . In Figure 8(a), the natural period ($\tau_o = 2\pi/\omega_o$) of the system is 1.014 times the duration of the pulse; in (b), 0.338 times the duration of the pulse; and in (c), 0.203 times the duration of the pulse. The damping ratio, δ_R, is indicated for each response curve.

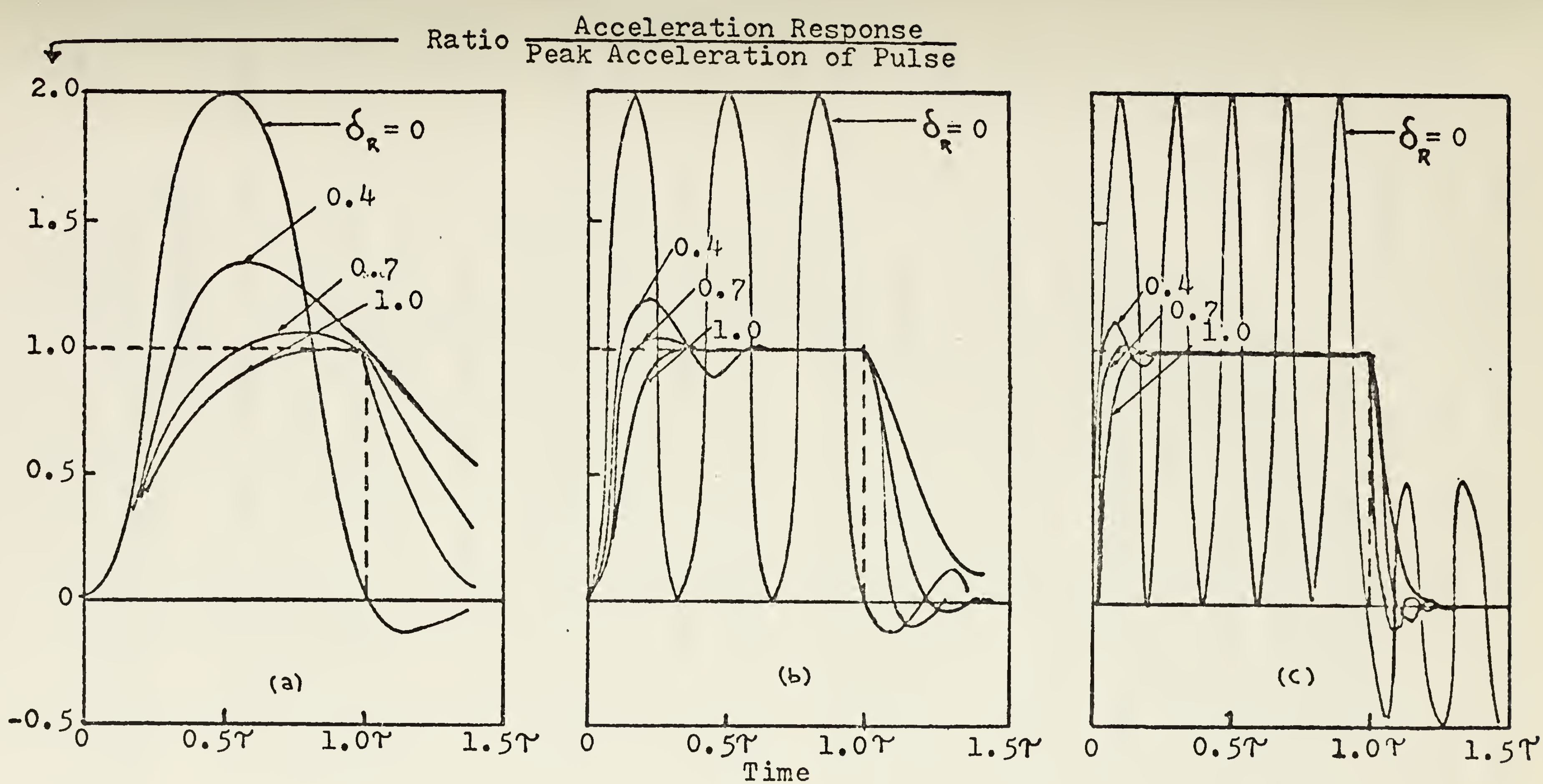


Figure 8. Acceleration response of an idealized seismic instrument to a rectangular pulse input transient.

Two observations can be made from the response curve shown in Figure 8. First, for moderate damping, the response of the system follows the pulse shape most faithfully when the natural period of the system is smallest compared to the pulse duration; i.e., when the natural frequency of the system is much greater than the fundamental frequency of the input pulse. Second, damping in the system reduces the response of the system at its own natural frequency; i.e., it reduces the transient vibration imposed upon the pulse. A study of the response of the system to other input wave shapes results in similar conclusions.

The steady-state response of the system, Figure 9, provides some additional information pertinent to the preceding observations.

It is readily seen that viscous damping of 0.6 to 0.7 of critical will result in a response that is very nearly constant with increasing frequency until the forcing frequency becomes equal to the undamped natural frequency of the system.

It may be concluded from the foregoing observations that if the system described represents an accelerometer, the device should have a high natural frequency and be operated below that frequency and have viscous damping of 0.6 to 0.7 of critical. Operation within these parameters will provide the best reproduction of the input shock pulse.

3. Accelerometers

There are several difficulties inherent in the design of an accelerometer. The major difficulties are:

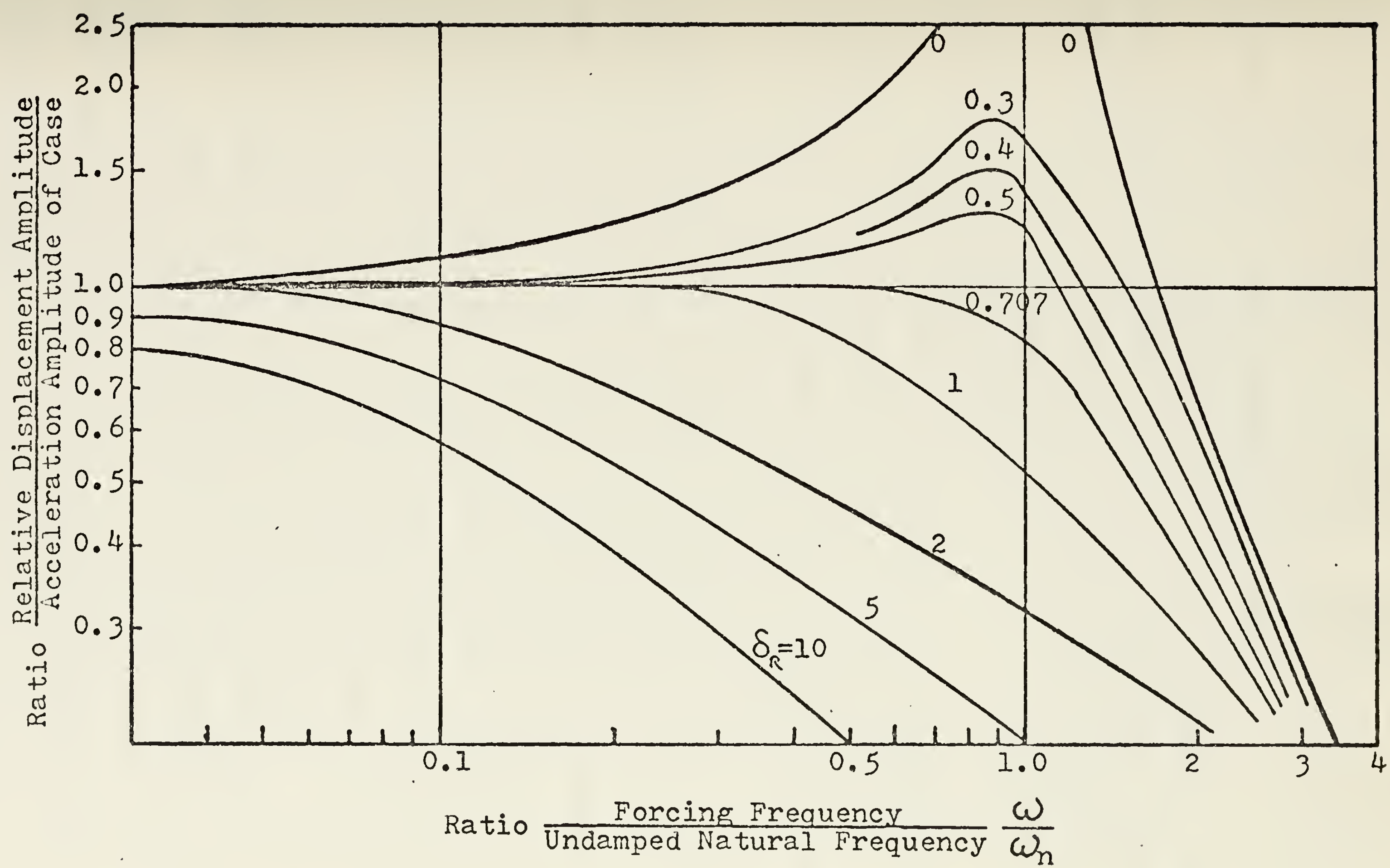


Figure 9. Steady-state response of an ideal accelerometer with viscous damping.

a. It is difficult to obtain purely viscous damping such as that described. Nonlinear damping will cause the accelerometer output to vary nonlinearly with the amplitude as well as with the frequency of motion.

b. Because of the high natural frequency required, the relative amplitude of motion is reduced and considerable amplification of accelerometer output may be required.

c. All accelerometers distort a complicated wave-shape because of phase shift.

Because of these difficulties it is seldom possible to obtain an accelerometer record of shock which is an exact reproduction of the actual acceleration pulse waveshape.

What, then, should be measured to best characterize an acceleration due to shock? There is no single answer. The parameters measured will depend upon the use for which data are required. If exact reproduction is required, the waveform of the input shock pulse could be measured and recorded by sampling at the Nyquist rate. Simpler schemes could be employed to measure peak acceleration amplitude and/or pulse duration. The peak acceleration amplitude is perhaps the most frequently measured parameter. The instruments of Table I are traditionally peak-measuring devices.

There are several types of accelerometers in general use. Two of the most common will be discussed briefly. A more thorough analysis of various accelerometer types is contained in Ref. 21.

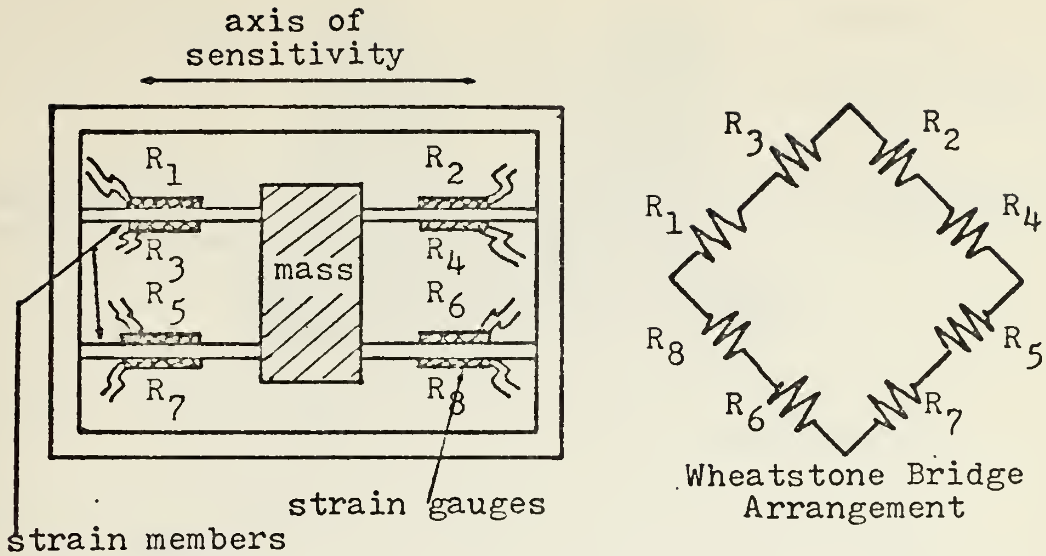


Figure 10. A bonded strain-gauge accelerometer.

a. The Strain-Gauge Accelerometer

In a strain-gauge accelerometer, the transducing element is usually composed of one or more fine wires arranged to be elongated or relaxed in response to the stress imposed on them. The change in length of the wire causes a change in its electrical resistance which is measured by an impedance bridge network. The transducer may be of the unbonded on bonded type depending upon how the wires are attached. A typical form of bonded strain-gauge accelerometer is shown in Figure 10. Strain-gauge elements other than wire are foil and thin films.

In contrast to the more commonly used piezoelectric accelerometers, strain-gauges are suitable for use at very low frequencies and offer good bidirectional output

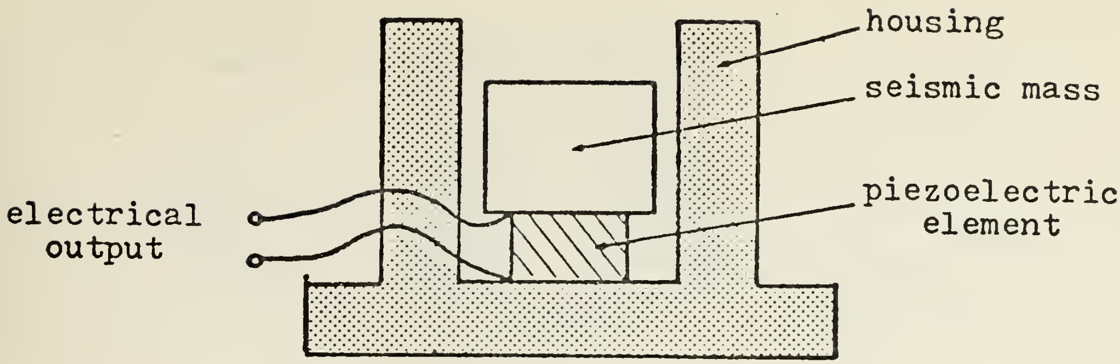


Figure 11. A simplified piezoelectric accelerometer.

linearity. However, the output signal level is lower than that of the piezoelectric type, typically about 50 millivolts maximum full-scale output for the unbonded type.

b. The Piezoelectric Accelerometer

Unlike the strain-gauge accelerometer, the piezoelectric accelerometer is a self-generating transducer and can be used with no external power source. It generates an electrical output signal proportional to acceleration but cannot be used at excitation frequencies near zero since little mechanical energy is being coupled into the system. This feature suggests that the strain-gauge accelerometer might be more useful for measuring steady-state vibrations of low frequency, but a piezoelectric type may show distinct advantages for measuring shock.

A simple piezoelectric accelerometer consists of a piezoelectric element similar to those in the section on hydrophones, but now the element is used to detect the acceleration of a seismic mass with relation to the accelerometer frame. A simplified representation is shown in Figure 11.

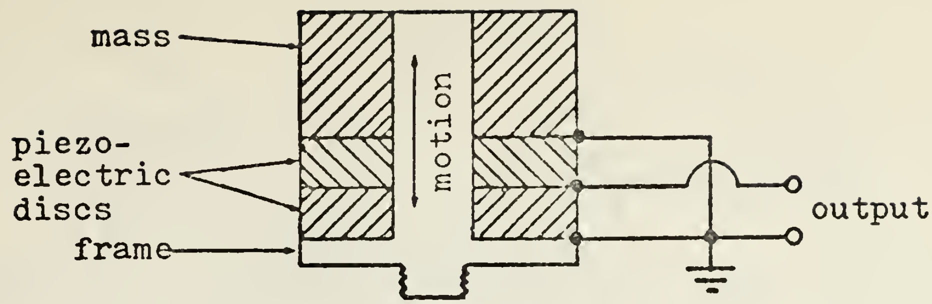


Figure 12. The compression-type accelerometer.

The mechanical equivalent circuit of Figure 7 applies. A thorough analysis of the piezoelectric effect is contained in Ref. 7 and will not be repeated here.

A piezoelectric accelerometer can operate in one of three modes: compression, shear, or bender. Each mode has certain advantages which make it suitable for a particular application.

An accelerometer operating in the compression mode, Figure 12, typically consists of a frame to which is fastened piezoelectric disc 1. This disc is provided with electrodes on upper and lower faces. A metallic disc is fastened to the upper face (for electrical contact), and piezoelectric disc 2 is bonded to the disc. Piezoelectric disc 2 is also electroded on upper and lower faces. Finally, a seismic mass is bonded to the upper face of disc 2. Acceleration along the axis of the assembly causes compressive or tensile forces on the piezoelectric element. Use of two discs makes it possible to connect the mass to the housing electrically so that insulators are not required. Since the element is poled along the axis and compression also

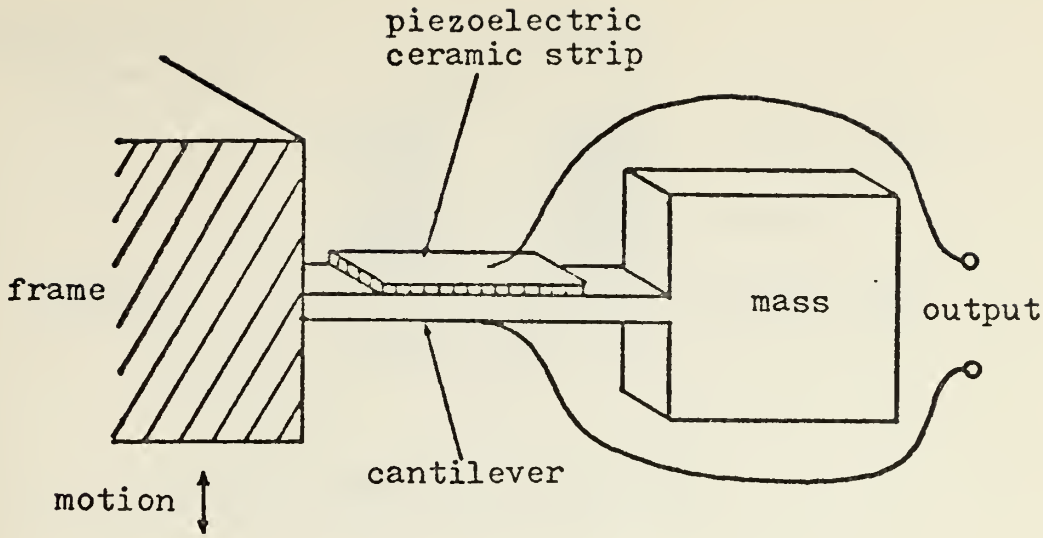


Figure 13. The bender-type accelerometer.

occurs along that axis, the electromechanical coupling factor of interest is k_{33}. The subscript 3 refers to the poling axis and 1 and 2 refer to arbitrarily chosen orthogonal axes in the plane normal to 3.

The bender-type accelerometer is shown in Figure 13. This device consists of a frame to which is attached a mass-loaded cantilever beam. Acceleration causes the cantilever to bend with a strain proportional to the applied acceleration. A strip of piezoelectric material with electrodes on top and bottom is bonded to the cantilever. Upward curvature causes compression in length of the piezoelectric element, and downward curvature causes tension. The resulting strain produces an output voltage proportional to acceleration. Poling is normal to the beam and in the direction of motion, but tension and compression occur parallel to the beam. Hence, the electromechanical coupling factor of interest is k_{31}.

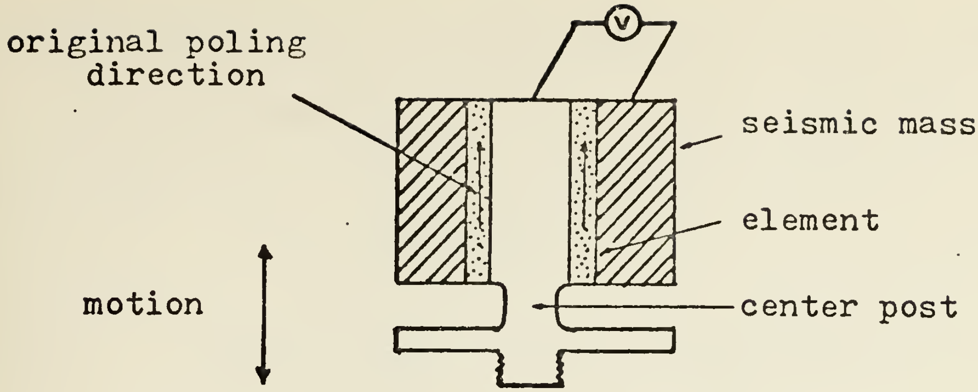


Figure 14. The shear-type accelerometer.

A typical form of the shear-mode accelerometer is shown in Figure 14. A tubular ceramic piezoelectric element with electrodes on inner and outer surfaces is bonded to a center post and a cylindrical seismic mass is bonded around the outer face of the ceramic. Accelerated motion along the axis of the cylinder causes the mass to exert a shear force on the ceramic element. The poling direction is along the axis and the generated electric field is normal to the poling direction. Hence, the electromechanical coupling factor of interest is k_{15}. Subscripts 4, 5, and 6 refer to shear stresses and strains in planes normal to the 1, 2, and 3 axes, respectively. In addition to the cylindrical design, shear-mode accelerometers can also utilize flat plates of ceramic to achieve the shear effect [24].

Lead zirconate titanate ceramics have become the dominant material for piezoelectric accelerometers. The piezoelectric constants, d_{ij}, and the electromechanical

coupling factors, k_{ij}, are well-known for various compositions of this material and permit a simple comparison of the three modes based upon the piezoelectric equation of state

$$D = dT + \varepsilon^T E \qquad (6)$$

Where D is the dielectric displacement vector,

T is the stress tensor,

ε^T is the electrical permittivity with constant stress,

and E is the applied electric field.

For a given accelerometer configuration, the dielectric displacement, D, can be related to the voltage which appears at the output terminals of the device.

TABLE II. Piezoelectric constants and electromechanical coupling factors for lead zirconate titanate.

k_{31}	0.31
k_{33}	0.67
k_{15}	0.69
d_{31}	-93.5×10^{-12} Coulombs/Newton
d_{33}	223 " "
d_{15}	494 " "

When there is no applied electric field, the equation of state (6) for both compression and bender types reduces to

$$D_3 = d_{31}(T_1 + T_2) + d_{33}T_3 \qquad (7)$$

or, substituting values from Table II,

$$D_3 = (-93.5(T_1 + T_2) + 223T_3) \times 10^{-12} \qquad (8)$$

Ideally, in the compression mode, $T_1 = T_2 = 0$. In the bender mode, $T_3 = 0$ and either T_1 or $T_2 = 0$. Clearly, for an equivalent amount of stress the compression-type accelerometer generates more than twice the dielectric displacement of the bender type. It is also evident that if the accelerometer is not ideally aligned with the direction of motion, spurious strains will occur in the transverse directions which will introduce error into the accelerometer output.

In the shear mode of operation, the equation of state (6) becomes

$$D_1 = d_{15}T_5 \qquad (9)$$

or,

$$D_1 = (494 \times 10^{-12})T_5 \ . \qquad (10)$$

If an equivalent stress is assumed, D_1, which is proportional to the voltage at the output terminals of the accelerometer, is more than twice that of the compression type and five times that of the bender type. Also, it is evident that the shear design appears independent of possible transverse coupling errors.

While it appears evident that the shear-mode design excels, it should be noted that each mode offers certain advantages.

The preferred design for most applications is bolted shear. It has the lowest sensitivity to such sources of error as base strain, thermal transients, acoustic excitation, and transverse vibration. The compression design is especially sensitive to these effects, but it gives the best combination of high output and high natural frequency. Bender and compression designs both have the advantage of lower cost than the shear mode design [32].

4. Conclusions

It has been learned that a shock or impact waveform can be measured accurately using an accelerometer. It is possible to measure shock amplitude, shock-pulse duration, or a close approximation of the exact wave shape of the shock pulse.

Since the shock pulse is typically of short duration hence rich in high frequency components, a piezoelectric accelerometer with its high natural resonant frequency is indicated. The choice of the piezoelectric accelerometer is doubly wise since its self-generating characteristic will reduce the power supply requirements of the portable recording device being considered.

The discussion of shock was restricted to single-degree-of-freedom systems; however, in reality it is not to be expected of the incident shock force. To determine the true magnitude of a shock pulse of unknown orientation, it will be necessary to mount three mutually perpendicular accelerometers and reconstruct the actual vector amplitude.

The use of the three accelerometers implies that low sensitivity to transverse vibration is especially desirable. For this reason, the obvious choice becomes a triaxially-mounted shear-mode piezoelectric accelerometer. One such device [32] has a resonant frequency of 30 kiloHertz and a sensitivity of 6 millivolts per g of acceleration. Amplification is required if the output signal is transmitted via connecting cables over a significant distance. The matching amplifier determines the low-frequency response of the system.

C. TEMPERATURE

1. Introduction

Although not specifically an acoustic quantity, environmental temperature is frequently of interest in the study of acoustic phenomena. The velocity of sound in a medium, for example, is a function of the temperature of that medium, and the temperature to which a shipping container is exposed is usually of great concern to the shipper.

The measurement of temperature can be accomplished using many different types of discrete sensors. Among these are the familiar glass-stem thermometer, bimetallic thermometers, filled systems (several types), resistance thermometers, pyrometers, thermistors, and thermocouples. Monolithic temperature-sensing devices have also been produced using integrated-circuit technology [8]. Of the many types of discrete devices, only a few lend themselves to electrical

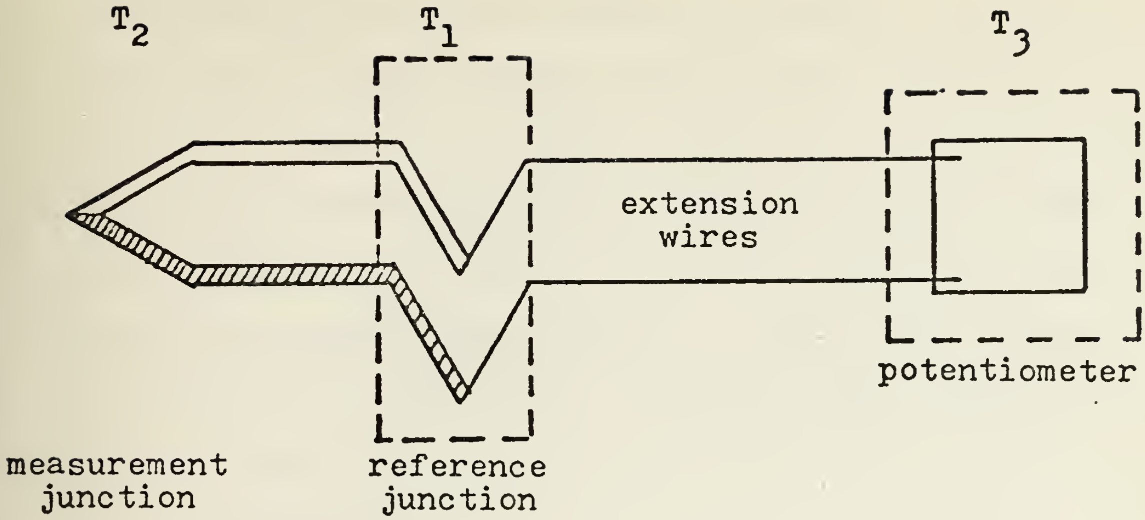


Figure 15. A practical circuit for thermocouple voltage measurement.

recording applications. Two of these, thermocouples and thermistors, and one monolithic device will be discussed.

2. Thermocouples

The thermocouple as a temperature-measuring instrument was first described by Becquerel about 1826 and, at that time, consisted of two bimetallic junctions held at constant but unequal temperatures between which a constant thermoelectric current was observed to exist. The voltage which produced the current was found to be determined by the temperature of the two bimetallic junctions and by the metallic composition of the thermoelements.

Practical thermocouple devices now generally incorporate a third material (the extension wire) which is at a third temperature, as shown in Figure 15 [26].

Thermocouples of the most frequently used types such as iron/constantan (an alloy of copper and 35-50% nickel) and

copper/constantan offer temperature ranges from -210°C to 980°C with ± 0.5°C accuracy, small size, lowest cost ($2-$50), ruggedness, and fast response. But as a class, thermocouples are susceptible to electrical noise, have small signal output and a sensitivity which is temperature dependent, have poor linearity, must be referenced to a known temperature, and require high gain amplifiers which may cost $200-$600 per channel [20].

3. Thermistors

Thermistors are essentially semiconductor devices which behave as resistors with a high temperature coefficient of resistance. In some extreme cases, the resistance of a thermistor may increase as much as 6% per °C rise in ambient temperature [39]. Over the range of -100°C to 400°C the temperature dependence is so great that there may be a change of 10 million to one in resistance. The result is an extremely high sensitivity (many times greater than that of a thermocouple) which makes the thermistor an unusually effective temperature transducer.

As opposed to the thermocouple, which generates a thermoelectric voltage output due to temperature difference, the thermistor dissipates electrical power while sensing temperature change. A typical dissipation constant in air is about 0.16 milliwatts/°C.[1]

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[1] Dissipation constant equals the power in milliwatts required to raise thermistor temperature 1°C measured with the thermistor suspended with its leads in the specified environment.

Thermistors have the significant advantage of providing direct temperature measurement where thermocouples provide only a relative indication based upon some reference temperature which might be difficult to maintain in a portable instrument.

Thermistors are inherently non-linear; however, this disadvantage may be compensated for by employing the thermistor in one leg of a bridge circuit to provide a linear temperature readout [39]. An optional method for providing linearity would be to store the inverse transfer characteristic in a digital memory and compensate for the non-linearity by signal processing.

The usable range of thermistors is more limited than that of thermocouples, about -100°C to 315°C with an accuracy of ± 0.24°C, and they are somewhat more expensive, about $10-$75.

A significant parameter of temperature sensors is the thermal time constant which is defined as the time required for the device to change 63.2% of the difference between initial and final temperatures when subjected to a step function change in temperature under zero-power conditions. In practical usage, the thermal time constant becomes a function of the mounting and circuitry as much as a function of the sensor itself. Responses of milliseconds to many minutes are attainable with both thermocouples and thermistors, but frequently time constants must be determined empirically before a final selection of the device can be made.

4. Monolithic Temperature Transducers

As pointed out in the preceding sections, both thermistors and thermocouples require external circuitry which might be expensive and wasteful in terms of space utilization and in terms of electrical power, particularly for applications in which space and power are of primary concern (e.g., portable instruments). An integrated-circuit temperature transducer has the potential for overcoming these drawbacks.

It has long been known that the mass-produced silicon transistor could serve as a transducer for general-purpose temperature measurements, but until recently such devices were more expensive than their thermistor or thermocouple counterparts. Recent improvements in integrated-circuit production techniques, however, have driven the cost of monolithic transistor temperature transducers below one dollar and made them price competitive.

The dual-transistor temperature-sensing element uses the difference in emitter-base voltage, ΔV_{be}, of two well-matched transistors operating at different current densities, I_{C1} and I_{C2}. The collector current ratio is controlled, so that the difference of emitter voltage is predictable and its manner of variation with temperature is well-known. A simple form of this device is shown in Figure 16. The output voltage, V_0, becomes a function of the temperature, T, and the ratios of circuit resistance; i.e.,

$$V_0 = \left(\frac{k}{q}\right)\left(1 + \frac{R_f}{R_b}\right) \ln\left(\frac{R_2}{R_1}\right) T \qquad (11)$$

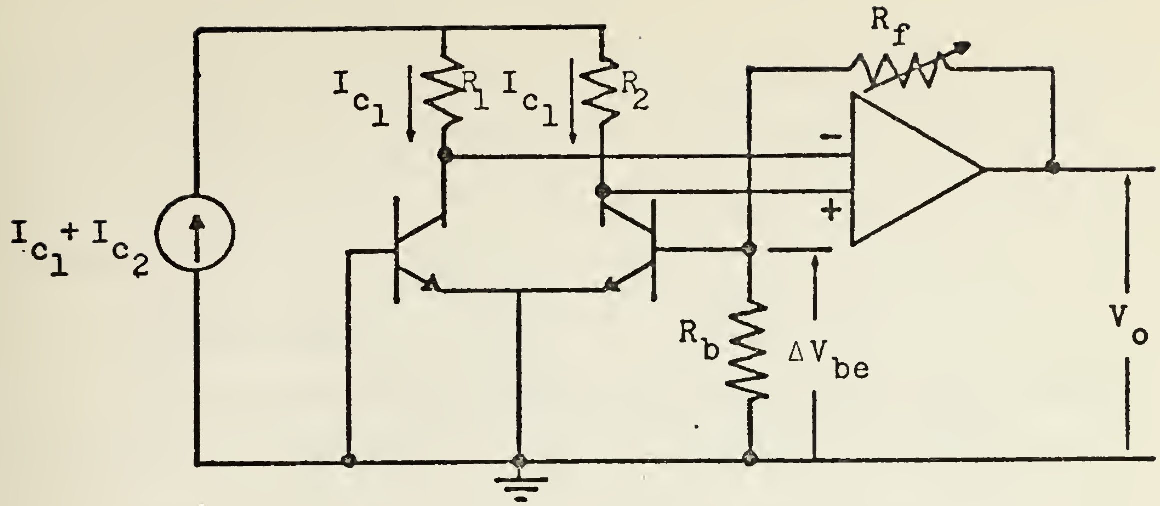

Figure 16. Dual transistor thermometer circuit diagram; the monolithic temperature transducer.

where k is Boltzmann's constant, and

q is one unit of elementary charge [42].

The integrated-circuit implementation of this device includes an operational amplifier with provisions for external feedback resistor, R_f, and also a Zener diode which, with the operational amplifier, forms a shunt regulator so that any temperature scale factor can be easily obtained with the addition of the proper value of external resistance.

The monolithic device described attains a linear output in the range -25°C to +85°C and can be calibrated to $\pm$ 0.5°C accuracy [8].

The simplicity, linearity, and low cost of the monolithic temperature transducer indicate that it is the best choice for a portable instrument. This device is packaged in several configurations to suit the needs of various

applications. Powered by any stable supply voltage higher than the required 6.8V reference, the maximum input bias current to the device is only 250 nano-Amperes. Output is a linear 10 mV/°C.

5. Conclusions

In environmental measurements, a relatively stable temperature is assumed. If the temperature of the ocean is to be measured, Duxbury [9] cites an annual range of ocean temperature of about -2°C to +30°C. A reasonable range for atmospheric measurements is about -30°C to +60°C. Thermistors, thermocouples, and monolithic sensors are all suitable for measurement of temperature in either medium, but the monolithic sensors appear to be most suitable for portable instrumentation.

D. RELATIVE HUMIDITY

Another parameter of interest in studying the dynamic environmental history of a shipping container is relative humidity. Relative humidity is defined to be the ratio of actual water vapor pressure in the air to the water vapor pressure in saturated air at the same temperature. The measurement of relative humidity partly justifies the simultaneous measurement of temperature since the percent relative humidity is meaningless unless the dry-bulb temperature is also known. Percent relative humidity can be obtained from wet-bulb/dry-bulb psychrometer data but is more commonly obtained directly using a Dunmore or Pope electric sensor.

Though not yet widely used, a solid-state sensor employing a polymer thin-film capacitor is also available for relative humidity measurement.

The Dunmore sensor employs a bifilar-wound inert wire grid on an insulative substrate which is coated with a lithium-chloride solution. The hygroscopic nature of lithium chloride causes it to absorb water vapor and change the ac resistance of the sensor. A typical range of sensitivity is 40 to 60 percent relative humidity.

The Pope cell employs a similar bifilar conductive grid, but it is mounted on a polystyrene insulative substrate which has been altered by sulfonation with sulfuric acid to make it hygroscopic. The presence of water vapor causes combination of sulfate ions in the substrate with hydrogen ions from the water vapor thereby altering the surface resistivity of the sensor as a function of humidity. The Pope cell is a wide-range sensor which is typically sensitive from 15 to 99 percent relative humidity.

In both Pope and Dunmore sensors, the element is arranged in an ac-excited Wheatstone bridge. Direct current polarizes the element and can cause a loss of calibration. The need for ac power is one of the drawbacks which make these two sensors undesirable for portable instrument use. Other drawbacks include limited ranges and slow response to changes in relative humidity--typically several seconds [45].

Solid-state technology has produced a new type of sensor which overcomes all of the faults of the Pope and Dunmore

sensors. The device is based upon the capacitance change of a polymer thin-film capacitor. A one-micron thick dielectric polymer layer absorbs water molecules through a thin metal electrode and causes capacitance change proportional to relative humidity. Response time is very fast, less than one second to 90 percent of the final value of relative humidity, and the sensitivity of the element is ± 0.5 percent relative humidity from 0 to 100 percent. The instrumentation amplifier which measures the capacitance change can be powered by low voltage direct current which makes it suitable for portable use [34].

Because of its suitability for portable instrumentation, its sensitivity, and its rapid response time, the polymer thin-film capacitor sensor is clearly the appropriate choice for an *in situ* recorder. The rapid response time may be especially significant if the device is to measure the relative humidity in the environment of a shipping container, since the statistics of such a situation cannot be easily predicted and rapid response may be necessary to detect all changes in relative humidity as they occur.

E. OTHER DATA

In addition to the physical data types discussed in the preceding sections, it should be noted that there are many other types of data for which *in situ* measurement and recording is appropriate.

One of these, hydrostatic pressure in the ocean, is related to ambient sea noise and was commented upon briefly in that section. At low frequencies, it is difficult to distinguish between the two effects. Hydrostatic pressure measuring devices are often piezoelectric transducers similar in principle to the hydrophones described earlier. Conceptually, measurement and recording of hydrostatic pressure are similar to measurement and recording of low-frequency ambient sea noise.

Likewise, seismic shock measurement is similar in concept to measurement of transportation shock. The transducers used for *in situ* seismic work are generally low-cost velocity pickups called geophones rather than accelerometers, but the data structure is much the same.

It is assumed that such data are truly unique only in the way they are sensed. That is, signal processing techniques which apply to the physical data types described thus far can also be applied to most other data types given the analog output of the measurement transducer. This assumption is suggested by classifying the data types in terms of frequency spectrum and periodicity as: 1) narrowband, either quasi-periodic (such as diurnal atmospheric-temperature variation) or aperiodic (such as relative humidity, salinity, etc.); or 2) broadband and generally aperiodic (such as transportation shock or seismic shock), although some vibration (such as ship vibration) could be described as periodic or quasi-periodic.

III. INFORMATION PROCESSING

A. INTRODUCTION

Thus far, the discussion has been limited to the sensing of data in analog form through the use of appropriate transducers. It is intended that this analog data will be stored in digital form in the solid-state memory of a recording device. To this end, the raw analog electrical output of each transducer must be processed and converted from its analog form to digital form. Processing will include amplification and other conditioning of the analog signal, analog-to-digital conversion, mathematical manipulation in digital form, and encoding for storage in memory.

All information processing must be accomplished with cognizance of the fact that a solid-state memory can be expected to have far less storage capacity than magnetic tape. Data must be densely packed to compensate for the reduced capacity and data reduction schemes become highly significant in the recorder design process.

The application for which the data is being collected determines what parameters of the physical quantity must be measured and how they may be processed and stored. For example, ambient noise could be recorded in such a way that exact reproduction was possible over a narrow band of frequencies, or in another case only the average amplitude level

of ambient sea noise might be recorded. Similarly, shock and vibration recordings might include full spectrum recordings or only peak shock amplitudes. The parameters of interest for temperature and relative humidity would normally be the amplitudes, but rate-of-change information could be extracted mathematically.

This chapter will first address the problems of analog-to-digital conversion in view of converting the four physical quantities of primary interest. Several types of A/D conversion and the advantages and disadvantages of each type will be discussed. The type of conversion device ultimately chosen will depend primarily upon the physical quantity parameters which are to be recorded and on the rate of data conversion and resolution required.

The remainder of the chapter will deal with the major problem of data reduction. Various methods will be discussed in detail based upon simple recorder system models and an appropriate data reduction scheme will be devised for the final system design.

B. ANALOG-TO-DIGITAL CONVERSION

1. Conversion Methods

A/D conversion methods can be classified into several types. In this discussion they will be classified into two groups: capacitor-charging types and discrete-voltage-comparison types. The capacitor-charging type depends upon digitally encoding the time required to charge a capacitor to some specified value, and the discrete-voltage-comparator type

achieves conversion by comparing the input voltage to fixed but discrete reference levels and generating an equivalent digital word whose value depends upon the results of the comparison. Important considerations in choosing the type of converter for a particular application will be the linearity of the device and the speed with which the conversion is accomplished. References 11, 22, 29, 26, and 46 describe the various A/D conversion techniques in detail.

2. Commercial Analog-to-Digital Converters

Commercial converters are classified as modular, hybrid, or monolithic. As the name implies, the modular device is an epoxy-encapsulated package usually less than the size of a human hand which contains all required circuitry and which constitutes the ultimate in state-of-the-art converter performance: 12-bit resolution in 2 microseconds. Hybrid devices require an external register and clock and typically have 20 kHz conversion rates. Monolithic devices are recent arrivals. These have the essential circuitry combined on one semiconductor chip. Only one on the market is complete. The others require an external reference or comparator. Conversion times are 18 to 40 microseconds [12].

C. DATA MANAGEMENT

It has been pointed out that analog or digital magnetic tape recording systems tend to make inefficient use of storage space by storing data in low-density form. For a solid-state recorder to be cost effective and space efficient,

data must be stored in high-density form by processing and encoding the input data in such a way that great data compression ratios can be attained. The examples which follow demonstrate the scope of this problem for ambient sea noise. Although not as well documented in the literature, shock data can be expected to require similar reduction prior to recording.

As a means of gauging performance, simple system models are presented in Section 2. The range of values specified for each parameter are incorporated into later designs. Based upon these models, several data reduction schemes are presented in order of increasing complexity and compared in terms of data compression ratios whenever such comparison is possible.

1. Some Examples of Ambient Noise Data-Recording Techniques

To reproduce an analog signal exactly, Shannon's sampling theorem must be observed. That is, the signal must be sampled at a rate at least twice that of the highest frequency in the input signal spectrum.

In many cases it is unnecessary to record the literal signal since only the average level of the signal must be known. Such cases permit signal processing which can greatly reduce the information rate and thus the required data storage capacity. The information rate depends very much upon the desired range of acoustic data required. A high-density recording system which collects data continuously from a wide

ambient-noise band has been described by Barbagelata [33], and other lower-density systems may obtain broadband data by recording for brief periods at predetermined intervals. Use of *in situ* signal processing to reduce the information rate has been infrequent.

Because of the variety of possible approaches to the problem of processing and recording ambient noise, three examples of existing systems will be discussed in terms of the goals of each system as well as the amount of data which must be stored in each. The examples to be discussed are:

a. Continuous recording of a 0-20 kHz band of ambient noise.

b. Continuous recording of low-frequency and infrasonic ambient noise.

c. Periodic sampling and recording of broadband ambient noise.

Reference 33 presents a high-density recording system for ambient noise. This system monitors ambient sea noise in the 0-20kHz band, filters the signal into several channels, converts the signal to digital form, and records the digital data on magnetic tape using high-density recording techniques. An estimate of the amount of data recorded is possible if some assumptions are made regarding the signal conditioning used.

Because of the extremely wide range of ambient noise amplitudes, logarithmic compression of the signal is advisable prior to A/D conversion to reduce the number of quantization

levels required. Wenz [44] has predicted a logarithmic range of approximately 70 decibels in the spectrum band of interest. This range can be described to $\pm$ 0.5 dB accuracy with 7-bit quantization and still provide latitude for excessively high levels. As noted, the sampling theorem requires 40,000 samples per second to reproduce the signal exactly.

From the above considerations one can calculate the amount of raw data in digitized form which must be recorded within a base period of 24 hours.

$$\text{Data Rate} = 40{,}000 \text{ samples/second} \times 7 \text{ bits/sample} \times 24 \text{ hours/day} \times 3600 \text{ seconds/hour} = 2.4 \times 10^{10} \text{ bits/day.}$$

Clearly, recording of broadband acoustic information for a prolonged period in digitized form requires a memory of formidable size. Magnetic tape systems can achieve such results since present technology has obtained densities of 6250 bytes/inch in typical tape drives, and the system described in Ref. 33 obtained a density of 12,000 bits/inch on each of 10 tracks.

In the case of infrasonic or low-frequency ambient noise the information rate is greatly reduced since the sampling theorem requires fewer samples per second. A typical recording system is the NUTMEG analog tape recorder [30] which directly records low frequency signals in the band from 20 to 300 Hertz. A digital version with 7-bit quantization

for encoding logarithmic amplitude would require 600 samples per second. Thus, in the same 24-hour period the amount of data obtained is:

$$\text{Data Rate} = 600 \text{ samples/second} \times 7 \text{ bits/second} \times 24 \text{ hours/day} \times 3600 \text{ seconds/day} = 3.6 \times 10^8 \text{ bits/day}$$

This data rate is two orders of magnitude less than before, but it is still quite large in relation to state-of-the-art solid state memory size which in one of the most dense configurations packs one million bits onto a circuit board of approximately 5x7 inches.

For the case of broadband ambient noise sampled periodically, the data rate can be further reduced. One such system developed by Naval Underwater Sound Laboratory, New London, Connecticut [31], for the purpose of correlating deep-ocean ambient noise spectra will be examined.

Broadband ambient-noise levels were recorded for two minutes every two hours. Twenty-five contiguous filter bands were used ranging from 11 Hz to 2816 Hz. Twelve samples per day were recorded on magnetic tape.

If it is assumed that the raw data was recorded in digitized form, the data rate can be estimated as before. Assuming data is sampled 5632 times each second for the duration of the 2-minute period which is repeated 12 times per day, and retaining the 7-bit quantization of amplitude, one can obtain the data rate for a recorder of this type.

$$\text{Data Rate} = 5632 \text{ sample/sec} \times 120 \text{ samples/sample period} \times 12 \text{ sample periods/day} \times 7 \text{ bits/sample} = 56.8 \times 10^6 \text{ bits/day}$$

This still seems to be a large number, but it is much more manageable than the data rate in the preceding examples.

In all the preceding examples, it should be noted that the spectrum recorded, the sampling period, and signal conditioning depend almost entirely upon the use for which the system is intended. For this reason, the examples serve only to indicate the amount of data involved when recording raw ambient noise in digitized form.

Little discussion was made of signal processing other than logarithmic compression of the amplitude levels. In any practical system; however, extensive processing could be used to reduce the amount of data which must be stored. For instance, an ambient noise sample of several minutes duration might be averaged in 1/3-octave increments and the band levels stored as 7-bit data words. A microprocessor-controlled solid-state recording device which could accomplish such an operation would be both small and inexpensive when compared on the basis of the size and cost of storing the raw data in the preceding examples. If only band levels are required from an ambient noise recording system, such a device is feasible, but if exact signal reproduction is required, such a device might be inappropriate depending on the bandwidth of the signal and the length of recording time required.

2. Elementary Recorder Models

As first simple models for the proposed solid-state environmental recorders, two separate devices are conceived. One will measure and record average ambient-noise levels as a function of time, and the second will measure and record the dynamic history of a shipping container, including peak shock amplitudes, environmental temperature, and relative humidity. Specifications for the devices will be limited to ranges of input values. Assignment of these values will permit numerical calculations which will aid in comparing various data reduction techniques.

The ambient-noise recorder will be modeled somewhat after the NUTMEG analog tape recorder of Ref. 30 which records ambient-noise continuously in the 20-300 Hertz band. But since continuous recording appears to be an expensive luxury in a solid-state device, average 1/3-octave band levels will be extracted and recorded. A dynamic range of 40 decibels will be assumed for deep-sea ambient noise in this band of frequencies.

The shipping-container recorder will be modeled after specifications for a similar device [41]. Peak shock amplitudes on three axes must be measured from zero to more than 120 g's. Only the greatest value must be recorded, but the direction of motion along the axis must be indicated by a sign to show acceleration or deceleration. Temperature must be recorded between -50°C and +65°C and relative humidity between zero and 100 percent.

3. Data Reduction

First considering the shock recorder, the data field for the indicated ranges of integer values might be presented in the decimal form

$$\pm \text{ SSS } \pm \text{ TT\%\%\%}$$

Where $\pm$ SSS is the subfield representing the signed integer value of greatest peak axial acceleration,

$\pm$ TT is the temperature subfield, and

%%% is the relative-humidity subfield.

If this data field is stored in binary-coded decimal form using a 4-bit digital word to represent each numeral or sign, as it might come from a digital voltmeter or a similar device, each field would require 40 bit-storage locations. In simple binary form each field would still require 31 bits. Either representation would be very wasteful of storage space since this data field can be significantly reduced. Several different data reduction schemes will be considered in order of increasing complexity.

a. Digital Encoding Using Compression Tables

The simplest form of digital data reduction is through the use of compression tables to reduce the data field size required. The compression can be further facilitated by quantifying the data into groups of levels or "windows" rather than recording the exact integer values. Digital encoding in this manner is routine since all analog-to-digital conversion requires some such scheme; however,

proper encoding can significantly reduce the size of the data field and the use of quantized windows can provide some increase in signal-to-noise ratio since the recorder becomes insensitive to small input transients.

Assume that shock is quantified into one of 21 possible levels as follows:

5 g steps from 0 to 80 g's

10 g steps from 80 to 120 g's

All shocks greather than 120 g's

Use of a sign bit to indicate acceleration or deceleration increases the number of levels to 42.

Temperature can be quantified into one of 23 discrete levels, each 5°C wide. No sign bit is required if the levels are referenced to -50°C. Relative humidity can be quantified into one of 10 discrete levels, each 10% wide.

A compression scheme can now be devised which uses a table of assigned codes for each level of the three physical quantities. For example,

Shock	Code	Temperature	Code
0-5 g	000001	-50 to -45°C	00001
5-10 g	000010	-45 to -40°C	00010
⋮	⋮	⋮	⋮
etc.			

The compression table for coding and decoding must reside in the data-management system of both the recorder and the reader.

Alternatives to the binary or BCD data field sizes can now be determined.

TABLE III. Alternative data-field sizes.

	Number of values, N	$\log_2 N$ Field size	Integer Field size
Shock	42	5.39	6
Temperature	23	4.52	5
Relative humidity	10	3.32	4
Total	75	13.24	15

The data field in binary coded decimal form required 40 bits. In compressed form the data field consists of only 15 bits, a compression ratio of 2.67:1. If fractional subfields are used, a compressed field of only 14 bits could be attained, but field extraction schemes for fractional subfields are extremely complicated and seldom worth the savings in space [1].

b. Delta Modulation

The preceding compression scheme can be, and usually is, applied to all data, but other techniques do not have such wide application. One such technique called delta modulation uses electronic circuitry to compare an analog input signal, m(t), with a generated stepwise approximation, $\tilde{m}(t)$. The name delta modulation derives from the fact that a signal proportional to the difference between the two is generated. The modulator output is a pulse train, $P_o(t)$. A typical set of waveforms are shown in Figure 17 with the digitally encoded output, $P_o(t)$. When the signal, m(t), exceeds the stepwise approximation, $\tilde{m}(t)$, a positive pulse is generated at the output, and when m(t) is less than $\tilde{m}(t)$, a

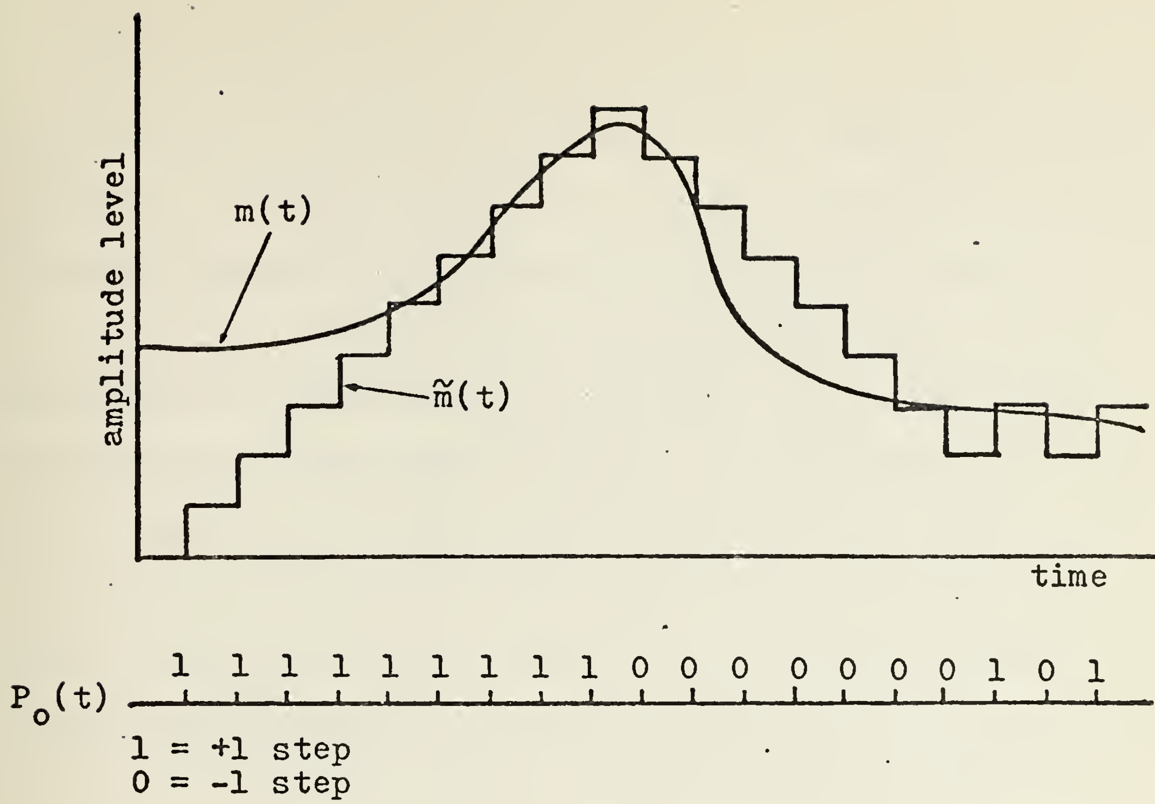

Figure 17. Delta modulation.

negative output pulse occurs. Delta modulation is essentially a version of the discrete voltage comparison A/D conversion method in which only the least significant bit is saved. Some quantization error (noise) is unavoidable as the approximation overshoots or undershoots the input, but a relatively good representation of the analog input is obtained when the input amplitude is slowly varying or constant, or when it has amplitude changes of less than the step size. However, it can be seen in the figure that when the signal changes rapidly, the modulator is overloaded because of the fixed step size in $\tilde{m}(t)$, and erroneous output occurs until the modulator can recover.

The point of slope overload occurs when the rates of rise of m(t) and $\tilde{m}(t)$ are equal. The sampling frequency required to prevent this error can be calculated by equating the two slopes if the frequency of the analog signal is known.

A modification of the delta modulator reduces the limitations imposed by both slope overload and input variations of less than step size. The adaptive delta modulator employs a variable gain amplifier which controls the step size. If the input variations are smaller than the step size so that the output consists of a string of pulses of alternating polarity, the integrated value of the output is zero. The modulator adapts to the zero integrated output by reducing the amplifier gain which reduces step size and increases the modulator sensitivity. If the analog signal starts to change faster than the modulator can follow so that the output becomes a string of pulses of the same polarity; the integrated output is a large voltage, the amplifier gain increases, and the step size also increases until a decrease in the integrated output voltage restores equilibrium [38].

The delta modulator and adaptive delta modulator differ in one other respect. While the reader device which decodes the recorded data has no difficulty decoding the ordinary delta modulated signal provided the step size is known, the reader for the adaptive delta modulator must have an adaptive adjustment capability resident in its decoder if the recorded pulse is to be properly interpreted.

Delta modulation appears to be most useful for analog signals of low frequency such as temperature or relative humidity. Another possible application is for encoding spectrum levels of deep-sea ambient noise if the raw noise data is first processed and averaged in analog form. It is unlikely that delta modulation would be used for encoding shock measurements due to the random statistics of shock data.

A dynamic range of 40 decibels was assumed for deep-sea ambient noise in the 20-300 Hertz spectrum band. Encoding this band requires a 6-bit data word for $\pm$ 0.5 dB accuracy. If delta modulation is used, only one bit of recorded data yields the same information. That single bit of data, however, has validity only with reference to the data which preceded it. The 6-bit encoded data word uniquely describes an ambient noise level. If the recorded data is to be read out serially, as for a strip-chart display, the delta modulation method is sufficient and represents a 6:1 compression of data. The comparison between delta modulation and the simply-encoded digital word can be stated another way: the simply-encoded digital word is <u>memoryless</u>, while the delta modulation method requires <u>memory</u> to reproduce the input signal with fidelity.

c. Predictive Encoding

A more sophisticated approach to data reduction is intermediate between the concepts of a fully-encoded (but memoryless) description of the analog input signal and the

delta-modulated description which requires a memory reference to previous values. In the predictive method, both the recorder encoder and the reader decoder estimate the current value of the analog input signal based on the previous values of the signal. The recorder encodes and stores the difference between the estimate and the true value. The subtracted portion is redundant and is not stored. The predictor may be linear (first order) or whatever higher order that gives the most satisfactory results.

The reader decodes the difference and reconstructs the true value using its copy of the estimated value. Since the difference is small compared to the true value of the signal, a considerable reduction in recorder memory size can be attained.

The necessary good predicted value of the input signal could be obtained for average deep-sea ambient noise level, temperature, or relative humidity, all of which are relatively stationary over short periods, based on the value of the signal at a fixed earlier period. For stationary signals, the predictor uses a fixed set of coefficients.

For non-stationary signals which are periodic or at least quasi-periodic, the predictor coefficients must be updated periodically to minimize mean-square error between the predicted and measured values. An adaptive system can be used to adjust the coefficients, but the new values must be stored for later use by the reader decoder. The additional stored values increase required memory size.

For aperiodic signals such as peak shocks, a predictive encoder is a poor choice since the current value of the input cannot be readily estimated from past history.

A choice between predictive methods or delta modulation must necessarily be a qualified one since no comparison of effectiveness exists in the literature. While perhaps more accurate than delta modulation, predictive methods requires more memory storage and more sophisticated computational methods in the data-management system.

d. Other Methods of Data Reduction

Elimination of redundancy in the analog input signal is the goal of a data reduction scheme. A significant reduction is generally obtained only at the expense of increased circuit complexity in the data management system.

The preceding section presented a device which predicted future response based on input data. For some physical quantities which have well-documented statistics, it should be possible to construct a predictive system which either stores a model of the known statistics or is capable of generating such a model.

The encoder of this generative system compares the analog input signal with the model and asks the question: is the value of the input signal the same as the predicted value? If the answer is "yes", all of the data in the input signal is redundant and nothing needs to be stored; if "no", only the deviation needs to be stored. So at worst, such a generative system could require more memory capacity than the

predictive system, but at best it could require little or no memory at all; that is, the statistical model and the input data could be in close correspondence. The amount of memory storage which the generative system would require would be based on the statistical variance of the model used.

The statistics of some types of ambient sea noise can be predicted (e.g., deep-sea noise), but generally with large variance ($\pm$ 5-10 dB). Diurnal temperature variation can also be modeled statistically for a fixed location. Even relative humidity may have some basis for statistical modeling which might be used for a generative method of data reduction; and for conventional shock or vibration measurements which store a replica of the input transient waveform, generative methods are also possible.

The predictive methods described previously are essentially generative methods. Linear prediction, as used in coding quasi-periodic waveforms such as speech, is somewhat more complex than the method described in the preceding section. The speech predictor uses a linear filter to obtain a weighted linear combination of the last N outputs of the filter and the present filter input. The linear combination serves as a predictor for the next filter output, and as in any predictive method, only the difference between the actual value and the predicted value must be stored. Reconstruction of the input is the function of the reader device.

It was seen in Figure 8 that an accelerometer analog output is quasi-periodic since it oscillates at its

damped natural frequency when excited by a shock transient. A linear predictor similar in concept to the speech linear predictor could be used to obtain a replica of the analog shock transient waveform. High sampling rates are essential for success with this method since the natural frequency of the accelerometer may be 20 kHz or higher. At least one sample every 25 microseconds must be assumed. Considerable speed is also required of the microprocessor since it must perform the mathematical operations which produce the linear combinations.

References 13 and 23 discuss a type of signal processing in the frequency domain called digital deconvolution which could prove useful for simplifying the shock transient waveform by removing distortions introduced by the motion of the foundation provided the frequency response of the foundation is known or can be determined. Digital deconvolution by fast Fourier transform is relatively straightforward but would require extremely high speed processing for real time applications. With existing technology this method is more suitable for the laboratory than for an *in situ* device.

D. CONCLUSIONS

The choice of method for data reduction in the cases of ambient noise, temperature, and relative humidity will depend not only on the level of data compression attained by the method but also on the computational speed which the selected method requires of the microprocessor-controlled data

management system. For the latter reason the following chapter discusses the state of the art in integrated-circuit technology.

IV. INTEGRATED-CIRCUIT ELECTRONICS TECHNOLOGY

A. INTRODUCTION

Before initiating designs for the proposed recorders, the state of the art in solid-state electronics must be investigated in order to select the most efficient hardware for the devices. In comparing the many different technologies, at least five important characteristics of each should be considered: speed, power consumption, density, cost, and availability. Operating speed is especially important in that it imposes limitations on the data-management system and on data reduction methods used by the system, but power consumption is also critical for the portable device proposed. Since different technologies operate at different voltage levels, level translation between devices must also be considered if more than one type of technology is to be used.

The two major semiconductor technologies are MOS and bipolar. MOS technology is based on the metal-oxide-silicon field-effect transistor (MOSFET) while bipolar technology is based on the ordinary epitaxial transistor. The two technologies can be viewed as general categories within each of which exists a number of different processes and techniques. These include PMOS (p-channel MOS), NMOS (n-channel MOS), CMOS (complementary MOS), TTL (transistor-transistor logic), and ECL (emitter-coupled logic), as well as the newer emerging

techniques of CMOS-on-sapphire and I^2L (integrated-injection logic).

A thorough description of these processes is far beyond the scope of this paper, but a brief discussion of the important characteristics will be presented to provide a basis for comparison. Reference 15 includes a concise and comparative operational description of the major processes to be discussed.

B. SEMICONDUCTOR TECHNOLOGIES

1. MOS Technology: PMOS versus NMOS

PMOS and NMOS integrated circuits are both based on the ordinary MOSFET. The former employs a P-type hole-conducting channel between source and drain, and the latter employs an N-type electron-conducting channel. Since electrons have a much higher carrier mobility than holes, NMOS gates have potentially higher switching speeds.

PMOS is the older MOS integrated-circuit process, but it offers few advantages over the newer NMOS process. PMOS excels in that more circuitry can be packed onto a smaller chip area, and PMOS has higher noise immunity than NMOS. But N-channel MOS is theoretically three times faster than PMOS, requires less power, and is much easier to interface with TTL. Since 1972, NMOS has been the MOS process offering the best combination of speed, density (4096 bits per chip), low power dissipation, and low cost (about 0.15¢ per chip in November 1975).

2. Complementary MOS (CMOS)

CMOS gets its name from the fact that both PMOS and NMOS devices are used together on the same chip as complementary pairs connected in series. When one device is conducting, at least one complementary device is held OFF so that the quiescent current is determined only by the leakage current of the OFF devices and is almost negligible. Power consumption effectively occurs only during switching times. For this reason, power consumption varies directly with operating frequency and is almost negligible when the device is in a standby mode. CMOS has high noise immunity and is operable over a wide range of temperatures. The cost (about 0.5¢ per bit) is greater than PMOS or NMOS since two devices are used where one would suffice, and for the same reason, the bit density per chip is also lower.

CMOS has a wide tolerance to power-supply voltage variations and can operate from an unregulated supply such as a battery. At higher supply voltages, switching speeds close to those of bipolar TTL can be achieved at the expense of greater power consumption. CMOS memories require higher supply voltages to obtain lower memory access times and thus cannot always benefit from CMOS low-supply-voltage advantages. Consequently, microprocessors and logic hardware are the primary applications for CMOS technology.

CMOS-on-sapphire is a newly emerging process. It is a variation of CMOS which attacks stray capacitance problems

inherent in MOS devices by isolating each unitary device in the chip on an insulating sapphire (aluminum oxide) substrate. With this technology speeds close to bipolar speeds can be achieved while still retaining the desirable low power features of CMOS.

At present, CMOS-on-sapphire is a relatively new technology and cannot supply a wide variety of devices. But when available, it provides the best combination of speed and low power although at the expense of bit density.

C. BIPOLAR TECHNOLOGY

Because of a more complicated cell structure, a bipolar technology produces optimum density at 1024 bits per chip, only one-fourth the density of MOS. Transistor-transistor logic (TTL) is based on the ordinary two-transistor multiple-emitter flip-flop with the ON transistors operated in saturation. It is several times as fast as MOS and requires only about a +5-volt power supply, but a TTL device consumes many times more power than an equivalent MOS device.

In another bipolar technology, emitter-coupled logic or ECL, on the ON transistor does not saturate. This eliminates the time required to bring the transistor out of saturation thus increasing speed, but the circuit still consumes considerable power. Unfortunately, ECL circuits require non-standard power supplies and are highly sensitive to changes in operating temperature. ECL is a poor logic choice unless ultra-high speed is the overriding parameter. No other

technology exceeds ECL on the basis of speed alone although one variation of TTL called Schottky TTL approaches ECL speeds.

Integrated injection logic (I^2L) is an emerging bipolar technology. It is basically a circuit-design technique which attacks the isolation requirements of bipolar devices by careful partitioning and by removal of unnecessary resistors. The technique significantly increases the density of random logic and appears to offer a combination of MOS density and bipolar speed; but presently, power consumption exceeds that of CMOS-on-sapphire which has much the same capabilities otherwise.

D. MEMORY TECHNOLOGIES

The MOS and bipolar technologies discussed so far are used in the design of integrated circuits ranging from simple random logic to more complex medium- and large-scale integrated circuits such as microprocessors and memory devices. Memory technology is not restricted to these processes; however, several other technologies exist which find application in memory devices alone.

It is useful to recall that several memory cell options are available to the designer. Each has been developed for a specific purpose.

The random access memory (RAM), as the name implies, permits storage of data at arbitrary locations in the memory. Data is written into the memory or read out under the address

control of the memory-management system. One special-use random access memory called a read-only memory (ROM) has a permanent or semi-permanent set of data written into it and is capable only of one-way transfer of data.

The alternative to the RAM is the serial-access design. This memory device requires sequential writing or reading of data. Magnetic tape is one example of a serial-access memory.

Memory devices can be further described as static or dynamic. Static cells are capable of retaining their data indefinitely. Dynamic memory cells must be refreshed periodically or the stored data is lost.

Integrated-circuit ROM and RAM designs can be implemented with either MOS or bipolar technology. Serial-access memories can use these processes as well, but other options include charge-coupled and bucket-brigade devices or magnetic-domain bubble devices, all of which are newly emerging dynamic memory technologies.

1. <u>Charge-Coupled Devices (CCD)</u>

Charge-coupled devices are basically dynamic shift registers and can be used wherever a serial-access memory is acceptable. Structurally, a CCD consists of 3 layers: an N-type or P-type substrate, then an oxide insulation layer, and a series of metalized electrodes over that. No P-N junctions are required, so that fabrication is greatly simplified and very cheap (about 0.1¢ per bit).

Information is stored in the CCD in the form of a charge. This charge is stored in a potential well created in

the semi-conductor substrate by the potential difference between adjacent electrodes. The charge is transferred along by creating an even greater potential difference with the next adjacent electrode. When the charge is transferred to the deeper depletion region which defines the newer potential well, the transfer voltages in the electrodes are removed and the charge is trapped in its new location. In this fashion data is recirculated continually through the CCD. One sequential location is available for reading or writing at the end of each transfer cycle.

Charge-coupled devices have achieved greater bit-density per chip (16,384 bits) than any other commercially available device and power consumption is only about one-twentieth that of MOS, but access times are much slower than for MOS or bipolar random access memories.

2. <u>Bucket-Brigade Devices (BBD)</u>

The bucket-brigade device is much like the CCD and has all of the CCD's advantages. The BBD uses a row of isolated-gate field-effect transistors with sources and drains connected in series and the gates capacitively coupled to the drains to effect the two-phase charge transfer.

As in the CCD, charge transfer in the bucket-brigade device is an inefficient process so that there is a definite ceiling on operating speeds that is many times slower than the speed of equivalent MOS or bipolar serial-memory devices.

Reference 43 describes both CCD and BBD in detail. While both processes have similar characteristics, CCD

memories presently dominate the market since they have proved less expensive to fabricate than BBD's.

3. <u>Magnetic-Domain Bubble Memories</u>

Still in the early stages of development, magnetic-domain bubble memories offer extremely high-density serial storage. Bell Telephone Laboratories has produced a 450-megabit memory in a package 9 by 5 by 2 centimeters and Rockwell has produced a 100-kilobit memory on a 300- by 300-mil chip [5]. But, to date, no commercial production has followed these laboratory efforts.

Magnetic bubbles are mobile cylindrical domains that can be generated in a thin layer of magnetic material and moved within the layer by interaction with set propagation paths processed onto the chip. Information is carried in the polarity of the bubble. A comprehensive description of magnetic bubble technology is contained in Ref. 5.

Magnetic-bubble devices are the slowest considered in this chapter with access times on the order of one millisecond. Although they have the greatest potential for high-density storage, the projected cost (about 0.05¢ per bit) is considered too high for commercial development at present. However, a recent fabrication process breakthrough by Phillips Research Laboratories in the Netherlands [10] may reverse this prognosis even though the new process yields lower bit densities and even greater access times than described above.

E. COMPARISON OF TECHNOLOGIES

On the basis of the preceding information it is possible to compare the various technologies in view of the weighted requirements of the proposed recorders.

For the microprocessor, random logic, and signal-conditioning devices required by the data management system, if cost is neglected and the assumption is made that device sizes will be acceptable, then the following evaluations will be made. Device technologies are listed in order of preference.

1. Low Power Consumption

Present	Future
a. CMOS (low power-supply voltages)	a. CMOS-on-sapphire
b. NMOS	b. CMOS
c. PMOS	c. I^2L
	d. NMOS

2. Speed and Power Consumption Equally Weighted

Present	Future
a. NMOS	a. CMOS-on-sapphire
b. CMOS	b. NMOS
c. PMOS	c. CMOS
d. TTL	d. I^2L

3. High Speed, Moderate-to-High Power Consumption

Present	Future
a. NMOS	a. CMOS-on-sapphire
b. CMOS (high supply voltages)	b. NMOS
c. TTL	c. I^2L
	d. CMOS

In some cases, the devices listed in the "future" column are commercially available now, but have not yet attained their maximum potential nor achieved satisfactory reliability. This evaluation is based on the near-term situation and must be reevaluated as new products appear on the commercial market.

Commercially available microprocessors vary from 4- to 16-bit wordlengths and are variously able to address from 4K to 64K words of memory. They include 15 PMOS microprocessors, 11 NMOS, 2 CMOS, and 8 bipolar, only one of which is I^2L [35].

For the memory device, high bit density becomes the major parameter for selection. Generally, the cost per bit is an inverse function of bit density so that cost can again be neglected. Memory devices are listed in order of preference.

1. <u>Moderate Speed, Lowest Power Consumption</u>

<u>RAM</u>	<u>Serial</u>
a. MOS	a. CCD (or BBD)
	b. MOS

2. <u>High Speed, Moderate Power Consumption</u>

a. Bipolar	a. Bipolar
b. MOS	b. MOS

F. BATTERY TECHNOLOGY

As a final word, something must be said about the power supply for the recording instrument. It is anticipated that the recorder will operate in its environment for at least two weeks. If the package design is to be small in size and

economical, the battery power supply must achieve high energy density, preferably at low cost.

Reference 27 describes state-of-the-art primary and secondary batteries. It is a concise comparison of their electrical characteristics and indicates that only two primary batteries approach the expected environmental temperature range requirements: the familiar alkaline cell and the newer lithium cell. Lithium cells appear to be more desirable since they offer more than 5 times the energy density of alkaline cells, as much as 150 watt-hours per pound. For example, a single 3-volt lithium cell the size of an ordinary zinc-carbon D-cell can power a flashlight continuously for 18 hours whereas the voltage-equivalent two zinc-carbon cells are exhausted after 15 minutes. Furthermore, the discharge curve of the lithium cell is flat so that a uniform voltage output is maintained throughout the useful lifetime of the battery. Voltage output of the cell is almost constant over the full operating range of temperatures, -50°C to +70°C.

Lithium cells suffer from some obstacles to production due largely to the tendency of lithium to react violently with water. But although developmental problems have resulted in high prices for the lithium cell, it is clearly the indicated choice for a long-term remotely-located instrument package. If lower power densities can be tolerated, low-cost alkaline cells should be used.

V. THE DESIGN OF THE RECORDER

A realistic appraisal of the preceding chapter indicates that the design options for a solid-state recording instrument are limited. Bubble memories might give the mass-storage capabilities of magnetic tape, but they are not yet available. The best combination of the other device technologies seems to be dependent upon the system specifications.

To illustrate the application of lessons learned in the preceding chapters, both the deep-sea ambient-noise level recorder and the shipping container shock recorder first specified in Chapter III.C.2 will be further specified and preliminary designs will be proposed in the remainder of this chapter.

A. THE AMBIENT NOISE RECORDER

1. Specifications

a. The recorder shall measure and record deep-sea ambient noise spectrum levels in the frequency range of 20 to 300 Hertz. A variance of ± 10 dB from Wenz's predicted 40-dB dynamic range will be expected. Resolution shall be 1 dB ± 0.5 dB.

b. Spectrum levels shall be measured in 1/3-octave filter bands and time averaged for 30 seconds. The average band levels shall be recorded. A time tag shall be recorded

each 10 minutes. Recording of data shall continue at the specified intervals for a period of two weeks.

c. The noise hydrophone is not specified, but should be similar to USRD H-58 which has a sensitivity of -94 dB re 1 volt per microbar over the band from 10 Hz to 4 kHz.

d. Frequency response of this system is unspecified but becomes a critical factor in the design of an actual device.

2. Design Considerations

Based on these specifications, eleven 1/3-octave bands can be described. Each must be averaged each 30 seconds. If eight analog samples in each band are averaged by the microprocessor during this period and if analog multiplexing and a single modular A/D converter are used, the aperture time for the converter can be as great as 340 milliseconds. Such a long period exceeds the aperture-time requirements of all commercially available modular A/D converters and permits the choice of slow-speed, low-power hardware. Lowest power consumption can be attained by using CMOS circuitry wherever possible except for a CCD serial memory and by using a combination of low power-supply voltages and low operating frequency. Logarithmic compression of the analog input signal and active filtering, which may be used to compensate for the non-linearity of the log device, may require non-CMOS hardware. All other devices are well represented in CMOS catalogs.

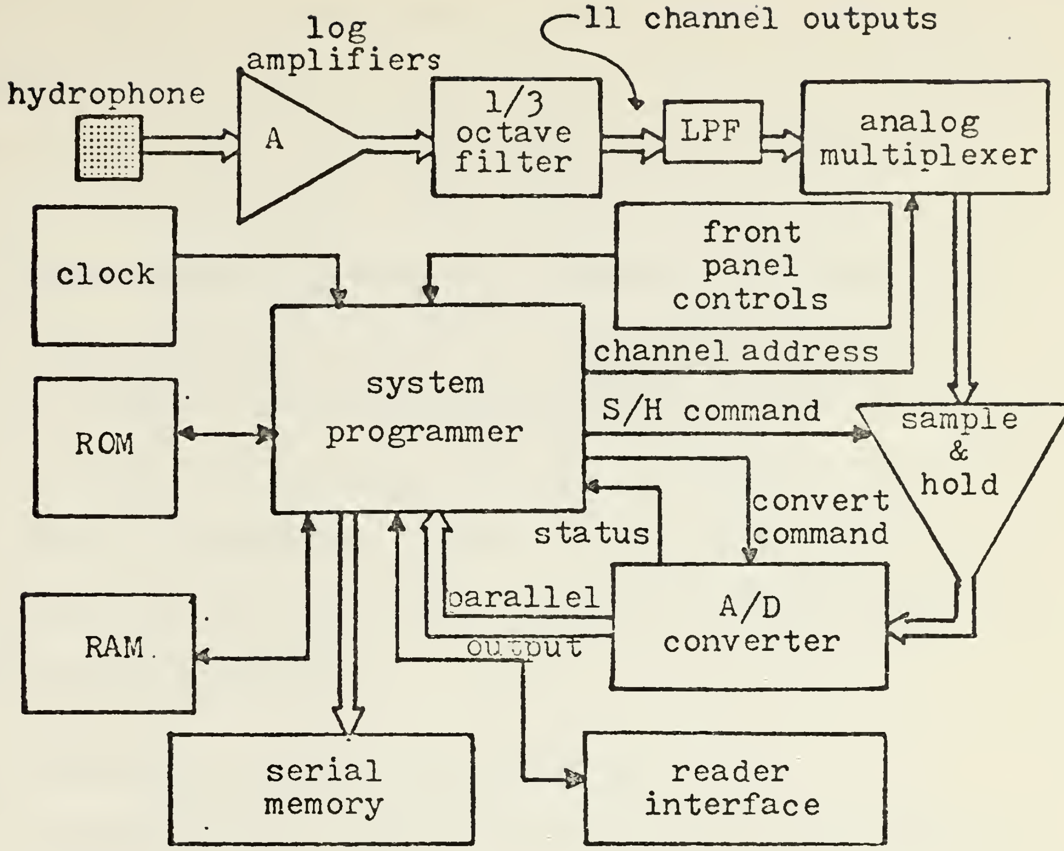

Figure 18. Ambient-noise recorder.

Figure 18 is a diagram of the proposed ambient noise recorder architecture.

The specified resolution of the spectrum levels requires 6-bit encoding. Over a period of two weeks, 241,920 bits of memory are required for the noise data. One time tag each ten minutes requires 2016 encoded time tags. Using 12-bit encoding, each time tag can be stored in two sequential locations. 24,192 additional bits of memory are required. The total mass-storage serial memory required is 266,112 bits organized into 6-bit words. The most straightforward version might use eighteen 16-kilobit CCD memory chips.

It should be noted that uniformly spaced time tags are redundant and need not be recorded. However, occasional recording of the time tag may be desirable to ensure valid data. If no tags are used, only 241,920 bits of memory are required but any data storage error which occurs could invalidate all subsequent data without alerting the user.

The machine language implementation of the controlling program for the ambient-noise recorder resides in a read-only memory (ROM) of unspecified length. Sixty-four words of random-access memory (RAM) are provided as a scratch pad for mathematical computation.

B. THE SHIPPING CONTAINER SHOCK RECORDER

U. S. Postal Service specifications [41] for a recording accelerometer are used as general guidelines for this device. These specifications are several orders of magnitude more complex than those of the preceding ambient noise recorder since the recorder must monitor not one physical quantity but three (shock, temperature, and relative humidity) and collect statistical data as well as a time history of events.

1. Specifications

a. Shock amplitude information shall be classified into discrete groups as follows:

16 groups of 5 g's each from 5 to 80 g's

4 groups of 10 g's each from 80 to 120 g's

1 group for all amplitudes exceeding 120 g's

b. The system shall record a time history of shocks occurring along three mutually orthogonal axes and distinguish

between acceleration and deceleration. The recorder shall resolve all events having a duration in excess of 4 milliseconds.

c. The system shall be capable of recording at least 65,000 events in each amplitude window without regard to the axis or direction of motion.

d. The system shall record temperature information in the range of -30 to +60°C. Temperature information shall be classified into nine 10°C amplitude windows. One temperature reading shall be recorded each minute and the system shall be capable of recording 65,000 events per amplitude window.

e. Once each minute the system shall record relative humidity in the range of 20 to 100 per cent. The information shall be classified into eight 10 percent amplitude windows. The system shall be capable of recording 65,000 events within each window.

f. The recorder shall be capable of continuous operation for a period of two weeks.

g. The recorder shall be capable of operating in one of four modes:

(1) Mode A. The maximum shock occurring during a 5-minute interval and its sign shall be recorded along with a time tag.

(2) Mode B. The same as mode A, but for a 30-second interval.

(3) Mode C. The same as mode A, but for a 3-second interval.

(4) Mode D. Standby operation until either of two conditions is met: expiration of 8 hours of standby operation or the beginning of container motion. When either condition is met, mode C is entered and routine operation begins.

2. Design Considerations

Based on the specifications, the worst-case mode of operation, i.e., the one which requires the most memory capacity, is mode C. Mode C will be used as a basis for calculations. It is assumed that the companion reader device will be capable of performing an inverse operation and recovering all data stored in the memory of the recorder. Since the reader device is not limited in either operating speed or power consumption, its design is less interesting and will not be discussed.

The specifications indicate the dual function of the recorder. On the one hand, a time history of events is recorded; on the other hand, statistical data, in the form of total number of events per amplitude window, is recorded for each physical quantity.

The design of the system hardware is straightforward. A sufficiently sensitive triaxial piezoelectric accelerometer operating in the shear mode (e.g., Gulton Industries TVA-3500) shall be used to obtain shock acceleration data along each of the three axes. Circuitry is provided to

amplify the voltage output of the accelerometers and to separate the output into 6 independent channels which contain either acceleration or deceleration information. Symbolically, x shall indicate the channel containing acceleration information along the x-axis, and $\bar{x}$ shall indicate the deceleration channel for the x-axis.

The six channel outputs are directed through a low-pass filter which detects the envelope of the shock waveform and reduces noise. Each channel signal then passes into a peak-amplitude detector which holds its information until analog-to-digital conversion. The peak detector is reset by a semiconductor switch after each conversion. This circuit and amplitude windowing account for all shock data compression other than digital encoding of the peak values. Rise time of the peak detector and low-pass filter as well as the frequency response of the accelerometer amplifier will satisfy the 4-millisecond signal-duration limitation imposed by the specifications.

The temperature shall be sensed by a monolithic semiconductor device (e.g., National Semiconductor Corporation's model LM 3911) of appropriate sensitivity. The output shall be amplified and low-pass filtered prior to analog-to-digital conversion.

Change in relative humidity shall be detected by a polymer thin-film capacitor sensor (e.g., WeatherMeasure Corporation's model HM-111P). The sensor output shall be amplified and low-pass filtered prior to A/D conversion.

Eight information channels arrive at the analog-to-digital converter. Performance is optimized by using one converter and analog multiplexing. In the worst case the channels must be sampled once each three seconds. But since temperature and relative humidity are only to be sampled once each minute, almost 500 milliseconds can be allotted to each of the shock channels during each three-second period. Such a long period places few restrictions on the choice of hardware.

Large-scale, low-speed, low-power consumption serial memory is clearly indicated for the time history data. The optimum choice is a charge-coupled device. A random-access memory with its shorter memory-access times seems appropriate for storage of statistical data, and since the only computations required consist of logical comparisons and additions, only a moderate-speed microprocessor is indicated. CMOS devices can be used for most modular components (e.g., for the microprocessor, Intersil's IM 6100; and for the A/D converter, Datel's DAS-16 LP series) as well as for all auxiliary circuit hardware, when available, to minimize power consumption.

Figure 19 is a diagram of the shipping-container recorder architecture. Once energized, operation of the recorder is continuous and storage of data is controlled by program interrupts.

The amount of memory required by the system can be approximated. For the statistical data: 6 shock channels of

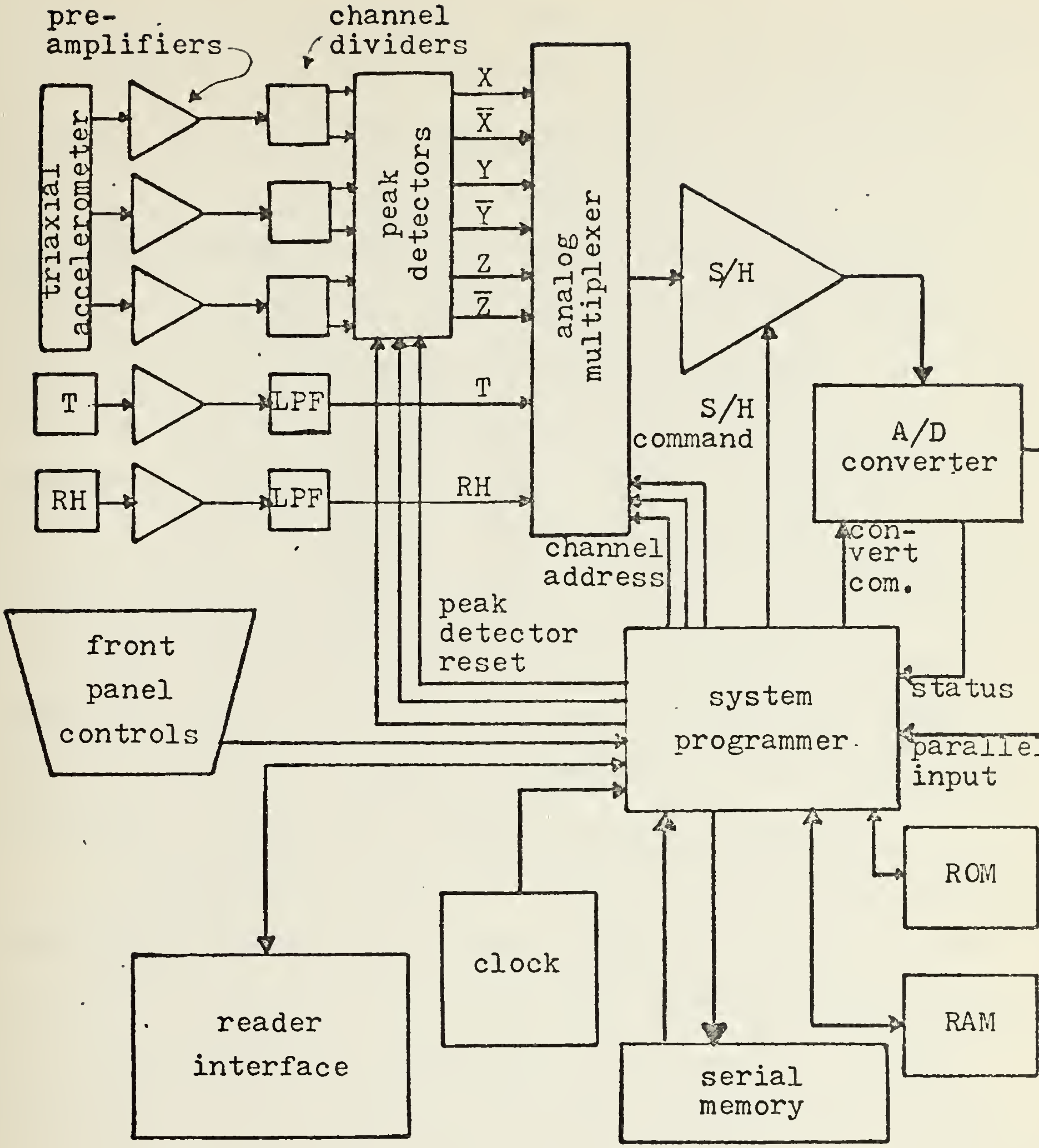

Figure 19. Shipping-container shock recorder.

21 amplitude windows must each record 65,000 ($\simeq 2^{16}$) events. 2016 bits of memory are required. Temperature requires 144 bits and relative humidity requires 128 bits for a total of 2288 bits of RAM.

For the time history data: the serial memory requires a data field of 19 bits to record a 3-second time tag discretely, but if a data-compression scheme is used which repeats a sequence of 1-minute time tags each hour, only 11 bits are required. The data fields for 21 signed levels of shock can be written in less than 11 bits as can composite data fields for temperature and relative humidity. 403,200 words are required for storage of shock data at 3-second intervals and 20,160 words for temperature and relative humidity at 1-minute intervals. Time tags require an additional 20,160 words for a total of 443,520 bits. If a microprocessor-compatible 12-bit wordlength is used, 5.4 million bits of CCD memory are required. Intel produces a complete CCD memory system arranged in 1-megabit segments on a 5- by 7-inch circuit board. System size appears to be no problem.

It should be noted that the system described is not completely efficient since some of the memory space is not utilized; i.e., 12 bits can represent 4096 different values, but is being used to represent not more than 21. A data-compression scheme using partial data fields could substantially reduce the size of the required memory. The preceding calculations are intended only to establish an upper bound on memory size and a practical system would surely require much less memory.

VI. CONCLUSION

This research has been concerned with several aspects of environmental data recording systems; i.e., the physical data to be recorded, the transducers required to sense the physical data, information processing of the data, and the hardware technology required to implement the system design. Considerations of physical data were limited to ambient sea noise, shock or impact, temperature, and relative humidity.

It was concluded that a piezoelectric hydrophone was best suited for the measurement of broadband ambient sea noise, due both to its uniform response below resonance and to its self-generating properties. Ambient noise spectrum levels were found to be more suitable measurement parameters than actual analog waveforms.

Methods of measuring shock or impact response of a fragile shipping container were considered in anticipation of designing a device capable of monitoring the dynamic behavior of the container and thereby gathering statistics or insuring that its contents were not exposed to a shock environment which exceeded shipping specifications. It was concluded that measuring and recording peak shock amplitude using a triaxial piezoelectric accelerometer operating in the shear mode is most feasible for portable devices. Great accuracy was not considered essential; thus measured values for the example were assigned to 5-g or 10-g amplitude windows.

Monolithic temperature sensors were deemed most suitable for measurement in the desired -25 to +85°C range, and relative humidity measurements limited to a range of 20% to 100% were found to be most easily obtained using a low-power polymer thin-film capacitor sensor.

An investigation of data reduction methods combined with an overview of solid-state technology led to the conclusion that CMOS-technology hardware with charge-coupled device memory offered the best combination of speed/power consumption characteristics for a portable environmental recorder and that the simplest data reduction methods consisted of logarithmic compression of the ambient noise data and windowing of shock, temperature, and relative humidity amplitudes. Other methods of data reduction considered were deemed inappropriate for a portable device at the current state of the art, due to the high speed/high power consumption requirements for the microprocesser and the short access times required for the memory of a high speed system.

The thesis of this research was to show the feasibility of replacing the ubiquitous analog magnetic-tape recorder with a solid-state environmental recorder for *in situ* acoustic measurements in hostile environments. It was shown by example that such devices are feasible for both shock measurements and ambient noise level recording.

It is concluded that a microprocessor-based solid-state recorder is practical if the expected use of the data permits signal processing to effect large-scale data reduction and

recording of the data in high-density form at low memory-access speeds. Recording of unprocessed raw data in digital form is not considered feasible, given the current state of the art in semiconductor memory technology, due to the size and projected cost of the required memory module and to the relatively long access times of large-scale memories. Real-time processing of a 10-kHz signal, for example, might require a new sample every 50 microseconds and access to the memory at the same rate. However, memory systems with the capacity to handle such data (charge-coupled devices or magnetic-domain bubble memories) have serial access times on the order of one millisecond--twenty times slower than required. The accessing problem is a major one for the designer and imposes great limitations on the use of solid-state memory.

Digital data acquisition systems, whether based on magnetic tape or on a solid-state recorder, have some inherent reliability problems which have not been resolved in this paper. Ambient noise digital recordings are particularly susceptible to acquisition of invalid data due to unusual noise sources. Analog magnetic tape recordings can be (and usually are) monitored while processing the recorded data, but one cannot listen to a digital recording. Spurious noise sources can go unnoticed and introduce false data into the recording. There is a real need for development of a signal-processing algorithm capable of discriminating against unusual or spurious noise sources before a truly effective solid-state digital recording system for ambient noise can be devised.

Failure of system components is another reliability area with which the designer must concern himself. The use of standard parts can greatly facilitate architectural design of the solid-state recorder, but parts failures have not been eliminated by current production techniques even though they have been significantly reduced. The CMOS integrated circuits which were found to be superior for portable instrument applications have the highest variations in reliability. One report places failure rates for CMOS devices as high as 0.5% per 1000 hours. It must be remembered that if an in situ recorder fails in use, the failure is undetected and the data of interest may be lost irretrievably.

It was noted in Chapter II that the shipping-container problem of distinguishing between container motion and vehicle motion has not been adequately solved. It is believed by the author that the digital deconvolution process offers a possible means of solving this problem and is a recommended area of future study.

The more sophisticated data reduction methods such as linear prediction or digital deconvolution all require high speed processing. It has been shown that high speed is obtained at the expense of high power consumption and is not usually feasible for portable instruments. However, if continuous operation of the device is not mandatory; that is, if the signal is such that a sustained period of input is followed by a long period with no input (e.g., a package in motion and then at rest); it might be possible to devise a

system which could operate at high speed during periods of input and then revert to a low power standby state during quiescent periods. The built-in control panel functions of some microprocessors suggest a means of implementing such a system.

It must be reiterated that the choice of system architecture will be dependent on the type of information the system must process. It may be that exact reproduction of the input analog waveform will yield significantly more information than a simpler peak- or average-amplitude recording device. If this is the case, then the extra expenditure for a high speed/high power consumption device is warranted. Further study is recommended in this area of system design.

Further effort is also recommended toward design of the solid-state systems, which were the subject of this research. The need for such devices is apparent. Magnetic-tape recorders really cannot compete with solid-state devices in terms of size, power consumption, resistance to mechanical shock, and signal-processing capabilities, but little work appears to be in progress toward commercial development of these devices.

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